[redacted]

FINAL ENGINEERING REPORT
FOR
NAVIGATION RECEIVER FOR USE IN RADUX LF NAVIGATION SYSTEM

297 155

This report covers the period 10 July 1956 to 1 July 1959

[redacted]

COMMUNICATIONS AND NAVIGATION LABORATORY
WESTERN MILITARY ELECTRONICS CENTER
MOTOROLA, INC
PHOENIX, ARIZONA

NAVY DEPARTMENT - BUREAU OF SHIPS - ELECTRONICS DIVISION

CONTRACT NO. NObsr 72578, INDEX NO. NE 010921

[redacted]

~~CONFIDENTIAL~~ UNCLASSIFIED

WMEC MOTOROLA PHOENIX
Control # MO/56-257
Copy # 7

Downgraded by BuShips ltr. Ser 685 B1a-391 1/30/59

FINAL ENGINEERING REPORT

FOR

NAVIGATION RECEIVER FOR USE IN RADUX LF NAVIGATION SYSTEM

Downgraded to unclassified by BuShips ltr. Ser 685 B1a-391 1/30/59

This report covers the period 10 July 1956 to 1 July 1959

~~WARNING: This document contains information affecting the national defense of the United States within the meaning of the Espionage Laws, Title 18, U.S.C., Sections 793 and 794. The transmission or the revelation of its contents in any manner to an unauthorized person is prohibited by law. Reproduction of this document in any form by other than activities of the Department of Defense and the Atomic Energy Commission is not authorized unless specifically approved by the Secretary of the Navy or the Chief of Naval Operations.~~

COMMUNICATIONS AND NAVIGATION LABORATORY
WESTERN MILITARY ELECTRONICS CENTER
MOTOROLA, INC.
PHOENIX, ARIZONA

NAVY DEPARTMENT - BUREAU OF SHIPS - ELECTRONICS DIVISION

CONTRACT NO. NObsr 72578, INDEX NO. NE 010921

~~CONFIDENTIAL~~ UNCLASSIFIED

DISTRIBUTION:

Dept. of the Navy Bureau of Ships	25
R. Hoffman	1
A. Hume	1
A. Grimaila	1
J. Kirch	4
H. Kelley	1

TABLE OF CONTENTS

TABLE OF CONTENTS (Contd)

TABLE OF CONTENTS (Contd)

LIST OF ILLUSTRATIONS

LIST OF ILLUSTRATIONS (Contd)

LIST OF TABLES

ABSTRACT

This is the final in a series of reports outlining the design, development, and fabrication of receivers for the Radux LF Navigation equipment for service test evaluation.

Part I, the General and Detailed Factual Data Sections, contains both a restatement of the purpose of the project and a detailed tabulation of the technical personnel effort expended.

The Detailed Factual Data Section comments on the design and development work accomplished on the Receiver Unit, Phase Comparator Unit, Power Supply Unit, Remote Monitor Unit, Remote Indicator Unit, Recorder Unit, Sequential Correlator, and Electromechanical data. These comments are followed by Test Results and a brief Conclusion.

Part II of this report is based on Recommendations for improvements in mechanical design and discriminator design.

PART I

PURPOSE, DATA, AND CONCLUSIONS

1. PURPOSE

This contract requires the design, development, and testing of two RADUX navigation receiver service test-models (fig. 1) that are to be evaluated for use in the RADUX LF Navigation System. Provisions of Specification SHIPS-N-2207, dated 3 January 1956, serve as design criteria for this development.

The equipment is to be used in a hyperbolic phase-comparison navigation system to measure the differences in arrival times of signals from three widely-spaced transmitters. The difference in times of arrivals from paired stations is proportional to the difference in the distances to each station and lines of constant differences in distance are hyperbolae. These lines of constant difference in distance are called lines of position, and the intersection of two such lines of position constitute a navigators fix.

The system is designed to give useful navigational data for ranges of 2000 to 3000 miles. The transmitted signals are 200-cps frequency modulation with a carrier frequency at or near 40 kc. The signals are phase locked and measurements are made on the phase of the modulation. Four separate transmitters, operating sequentially, can be accommodated. The total transmission time is 3 seconds with each transmitter operating for a different period of time.

The primary function of the receiver is to determine the difference in phase between the pairs of transmitted signals. To accomplish this, the receiver must first properly identify each of the transmitted signals and then compare the phase of each of the transmitted signals to a locally generated stable signal. The difference between two such compared signals is the desired line of position.

Identification of the particular received signal requires that the receiver generate a signal sequence that is identical to the transmitted sequence. For RADUX, this sequence is 7, 8, 9, and 10 units with each transmission unit separated by 2 off units. Since each unit is 60 milliseconds long, the total transmission interval is 3 seconds. The locally generated signal sequence is compared to the incoming signal sequence in a series of coincidence gates, and the local sequence is advanced or retarded automatically as required. Once synchronized, the local gates are used to switch the incoming signals to the individual measuring channels at the proper time.

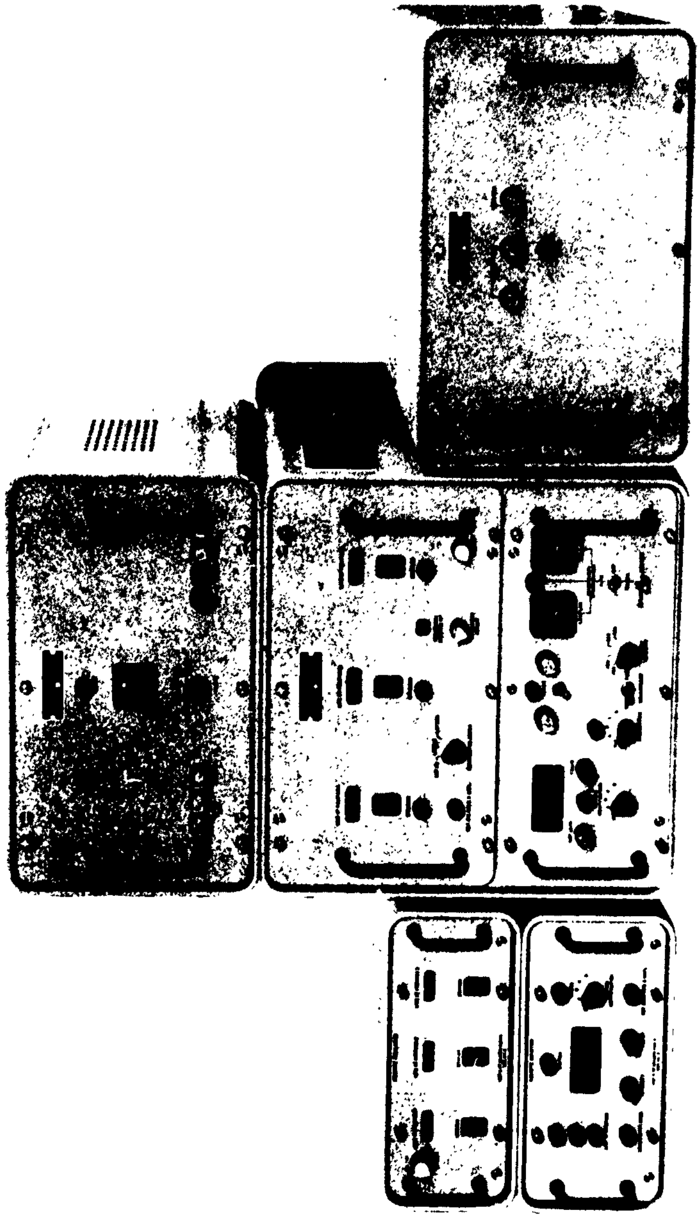


Figure 1. RADUX Navigation Receiver Service Test-Models

RADUX is a precision instrument and in order to provide accurate and reliable navigational information, the equipment must be capable of making precise measurements under extremely poor signal to noise ratios and under conditions where sequential signals can vary by as much as 100 db.

An important application of this equipment is for submarine operation under submerged conditions. For this application, reception would be from a crossed loop antenna with extremely small signals present. This requires that the receiver circuits be highly stable and sensitive and at the same time not introduce excessive phase shifts with signal level changes or loop orientation. Both of these objectives are attained through the use of a sharply tuned antenna input circuit followed by an operational amplifier that serves to flatten the phase response characteristics. Provision is also made for a whip antenna input for general purpose and shipboard use.

1.1 DESIGN OBJECTIVES

1.1.1 Receiver

The receiver shall contain one superheterodyne channel capable of receiving frequency-modulated signals on frequencies of 34, 36, 38, or 40 kc, with a constant receiver phase delay. Sufficient amplification shall be provided to saturate the limiters with an input of one microvolt or less in an impedance of 50 ohms to a standard dummy antenna. A discriminator shall be provided to extract the 200-cps modulation.

The 3-db bandwidth shall be 600 ± 60 cycles with at least 60-db attenuation at or more than 2000 cycles away from the center frequency. Normal operation shall be obtained with an input voltage differential up to 100 db with a phase shift not exceeding 5 microseconds. Maximum signal input shall be 300,000 microvolts.

The loop-antenna coupling circuits in the receiver unit shall meet the following requirements:

(a) Sensitivity measurements shall be made using a shielded booth and 500-ohm transmission line with a line constant of 7.75, energized by a standard signal generator over the specified frequency range of 34, 36, 38, and 40 kc. The sensitivity on omnidirectional operation shall be 200 microvolts, or less, as indicated by the signal generator output with the transmission line connected to obtain a 20-db signal-to-noise rms voltage ratio measured in the 600-cycle receiver bandwidth. The sensitivity on single plane operation shall be 150 microvolts or less to provide the same output. The deviation of

the omnidirectional phase pattern from a uniform pattern shall not exceed 2 microseconds at any of the specified frequencies. Means shall be provided for selecting single-plane loop operation on either of the crossed loops or for selecting omnidirectional loop operation.

(b) Using a RADUX system signal generator connected to the transmission line, a carrier input of 1000 microvolts for each commutated signal shall be applied and the average line-of-position reading noted for each line of position. When the carrier of one signal is reduced in 6-db steps so that the signal-to-noise ratio in the receiver bandwidth is one-to-one, the change in the mean of the associated line of position shall not be more than 5 microseconds. A carrier input of 500,000 microvolts for each signal shall be applied and the average line-of-position readings noted. When the carrier of one signal is reduced to 50 microvolts in 6-db steps, the change in the mean of the associated line of position shall not be more than 5 microseconds. These requirements apply to both single-plane and omnidirectional operation over the specified frequency range.

1.1.2 Phase Comparator Unit

The navigation-receiver indicator shall be capable of measuring the phase difference between the 200-cycle modulation on signals from a RADUX system signal generator to the accuracy specified under any of the conditions listed in items (a) through (e). The measurements shall be made on signals with modulation indices of $\pi/4$ (90 degrees total r-f carrier shift). White noise levels shall be equivalent to those obtained from a standard General Radio type 1390A random noise generator in a noise bandwidth of 600 cycles to the 3-db points. The signal and noise voltages shall be mixed in a circuit with a 50-ohm output impedance. The mixed signals shall be coupled to the antenna coupler for the whip antenna through a standard dummy antenna.

(a) With 10 microvolts carrier signal and 20 microvolts rms white noise for one commutated signal and with 200 microvolts and no noise for another commutated signal, the standard deviation of 120 time-difference readings taken at 30-second intervals for a 1-hour period shall not exceed 10 microseconds.

(b) With 1 microvolt signal input for all four signals and only internally generated noise, the standard deviation of 240 readings on each line of position, taken at 6-minute intervals for a 24-hour period, shall not exceed 5 microseconds.

(c) With 10 microvolts signal input for each signal, the receiver indicator shall follow a 30-microsecond instantaneous change in phase of one signal such that at least 70 percent of the change shall be corrected in one minute.

(d) With a signal input of 10 microvolts for one signal and 200 microvolts for another, the change in the mean indicated-time-difference readings shall not exceed 5 microseconds when a white-noise signal of 20 microvolts is added to the input signals.

(e) Under conditions simulating the receiver indicator in motion along the baseline of two transmitters at a 40-knot rate, the mean lag error in the indicated phase difference between typical pulsed input signals of 10 microvolts, with a noise input of 10 microvolts, shall not exceed 10 microseconds.

A coherent r-f phase detector shall be incorporated in this unit to distinguish between signal and noise in the i-f output of the receiver unit. The output of this circuit shall be applied to a cross-correlating commutation synchronizer. This synchronizer shall provide an error-voltage output to precess an electronic commutator bringing it into coincidence with the signal sequence pattern when any two of the four possible signals are received. The time spent in synchronizing the commutation shall not exceed 3 minutes when the signal-to-noise ratio in the receiver bandwidth exceeds 1 to 5. The total time required for synchronization and phase matching of the receiver shall not exceed 15 minutes, starting from a cold start (i.e., with the equipment warmed up only).

2. GENERAL FACTUAL DATA

2.1 IDENTIFICATION OF PERSONNEL

Technical personnel whose efforts were expended on the problem are listed in Table 1.

TABLE 1

TECHNICAL PERSONNEL EFFORT

Name	Title	Hours
James Kirch	Project Engineer	2094
Willis Steen	Project Engineer	1784
Bernard Parmet	Staff Engineer	341
George Monser	Staff Engineer	304
Morton Stern	Electrical Engineer	1734
Robert Hansen	Electrical Engineer	1066
Warren Palmer	Electrical Engineer	3969
Stanley Malinowski	Electrical Engineer	573
John Corkill	Electrical Engineer	911
Frederic Harris	Electrical Engineer	1809
Jack Niebell	Electrical Engineer	424
John Barto	Electrical Engineer	2556
John Knudsen	Mechanical Engineer	1064
David Mocsny	Mechanical Engineer	331
Theodore Kappel	Engineering Ass't	2977
Earl Willis	Technician	3507
Ronald Petz	Technician	1293
Steve Wronski	Technician	1064
Rudy Brabec	Technician	504
Lawrence Tonies	Technician	144
Charles Hurt	Technician	382
Patrick Galbraith	Technician	2408
Richard Netzel	Mechanical Designer	2970
Arthur Burton	Mechanical Designer	1602
Ronald Hawkins	Mechanical Designer	777

2.2 PATENTS

A method of obtaining maximum available sensitivity from an antenna with a minimum transfer phase slope has been devised by Warren Palmer and Bernard Parmet. The new method eliminates the need for a large, high-loss input transformer by replacing it with a light-weight, low-loss, non-critical circuit with properties approaching those of an ideal transformer while minimizing coupling losses from the antenna to the receiver. An operational feedback amplifier is used in conjunction with a tuned antenna to realize these properties.

Patent application presently is being processed by Mueller and Aichele, 105 West Adams Street, Chicago, Illinois.

2.3 REFERENCES

The following is a list of references cited in portions of this report. The order in which each reference appears in this list does not necessarily reflect the chronological order of appearance in the text but rather acquaints the reader with the reference titles. Footnotes are used to indicate the particular reference cited in the text.

Argiumbau, L. B., Vacuum Tube Circuits, New York: John Wiley & Sons, Inc., 1948.

Chestnut, Harold and R. W. Mayer, Servomechanisms and Regulating System Design, New York: John Wiley & Sons, Inc., 1952.

Ertman, R. J., Investigation of Underwater Radio Reception, Naval Research Laboratories, Washington D.C., Report No. 4406, 1954.

Gray, Truman S., Applied Electronics, New York: John Wiley & Sons, Inc., 1954.

Harrington, J. V., "An Analysis of the Detection of Repeated Signals in Noise by Binary Integration", IRE Transactions on Information Theory, March, 1955, Vol. II-1.

Interim Engineering Report No. 18 For Navigation Receiver For Use In RADUX LF Navigation System, Motorola, Inc., Contract No. NObsr 72578, March, 1959 (Confidential).

Reference Data for Radio Engineers, Fourth Edition, New York: International Telephone and Telegraph Corp., 1956.

Report On Noise In FM Systems and Loop Sensitivity In RADUX System, Sperry Gyroscope Co., Publication No. 5221-1399, October, 1955 (Confidential).

Sturley, K. R., Radio Receiver Design, Second Edition, New York: John Wiley & Sons, Inc., 1953.

Truxal, John G., Control System Synthesis, New York: McGraw-Hill Book Company, Inc., 1955.

2.4 MEASUREMENT PROCEDURES

The following subparagraphs describe the various test procedures performed on the RADUX navigation receivers. The only test equipments required are a General Radio type 1390A random noise generator and two Hewlett-Packard type 400C vacuum-tube voltmeters.

2.4.1 Shift In Mean (fig. 2)

The simulator supplies 10 microvolts of signal during the A and C transmission periods and 200 microvolts of signal during the B and D transmission periods. The STATION SELECTOR switch on the RADUX phase comparator is placed in the ABC position. The RADUX equipment is synchronized and a 15-minute period is used to permit the phase indicator system to null completely. The mean reading (X_o) is taken after the 15-minute null period.

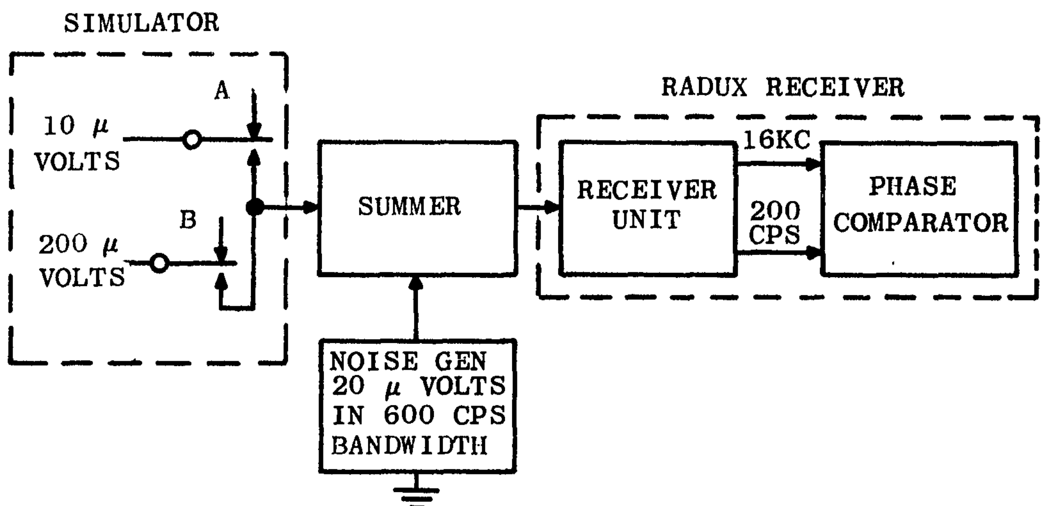


Figure 2. Shift In Mean Reading Test, Block Diagram

The noise generator is then turned on with the amplitude (previously adjusted) producing a 20-microvolt input in a 600-cps bandwidth. Readings are taken for 75 minutes at 1/2-minute intervals. The first 15-minutes' readings are not used in the calculations; however, the last 60-minutes' readings are added together and divided by the total number of readings (120 in all) to find the average mean reading ($\bar{X}_o$).

The shift in mean reading (X_s) is the difference between the average mean reading with noise ($\bar{X}_o$) and the original mean reading before noise was added (X_o), i.e.,

$$X_s = \bar{X}_o - X_o \qquad (1)$$

2.4.2 Standard Deviation

The standard deviation (σ) is calculated from the following formula, using the same 60-minutes' readings used to calculate

the mean reading:

$$\sigma = \sqrt{\sum_{i=1}^{n} \frac{(X_i - \bar{X}_o)^2}{n}} \qquad (2)$$

where:

X_i = any single reading and

n = total No. of readings.

2.4.3 Standard Deviation With One-Microvolt Input (fig. 3)

The simulator supplies a 1-microvolt signal to the RADUX equipment with only receiver noise present. The RADUX equipment is synchronized and a 15-minute waiting period is allowed to stabilize the initial phase reading. Readings are then taken at 6-minute intervals for each line-of-position over a 24-hour period (240 readings in all). The standard deviation for each line-of-position is calculated from equation (2) in respect to the mean reading over the 24-hour period, i.e., n = 240.

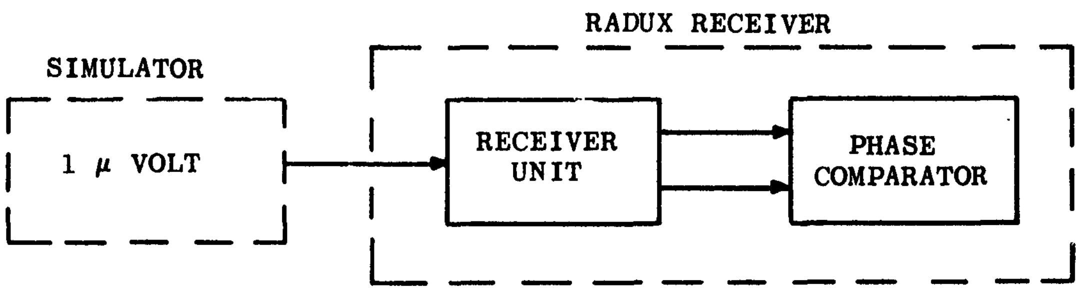


Figure 3. Standard Deviation With One-Microvolt Input Test, Block Diagram

2.4.4 Transient Response (fig. 4)

The simulator supplies 10-microvolt signals of one phase during the A and C transmission periods and another phase during the B and D periods. (The phase of each of the two signals is independent of the other.) The STATION SELECTOR switch is placed in the ABC position, the RADUX equipment is synchronized and 15 minutes is allowed for the phase indicator to indicate the phase-difference readings.

The phase of the 200-cps modulated signals is changed rapidly by 30 microseconds and the reading of line-of-position No. 1 (A-B) is recorded at 10-second intervals for 2 minutes. The time required by the phase indicators to correct 70 percent of the phase change (21 microseconds) is the test result.

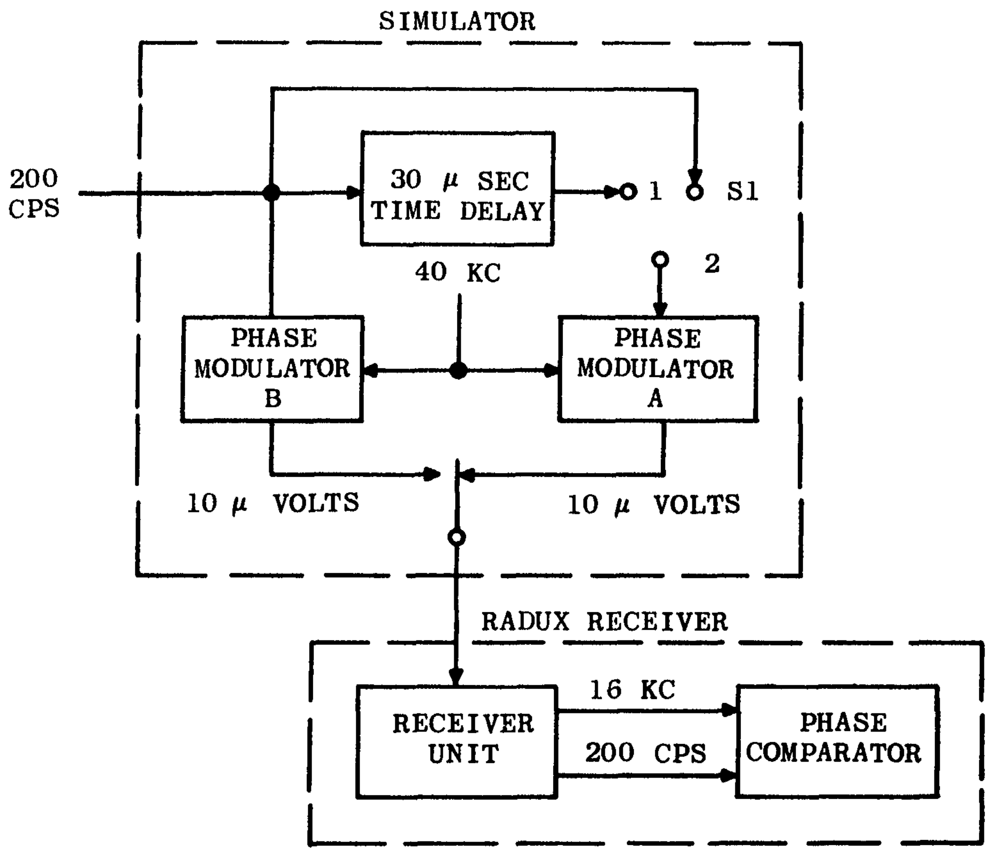


Figure 4. Transient Response Test, Block Diagram

2.4.5 <u>Tracking Test</u> (fig. 5)

The simulator supplies 10-microvolt signals of one phase during the A and C transmission periods and another phase during the B and D periods. (The phase of each of the two signals is independent of the other.) The STATION SELECTOR switch is placed in the ABC position, the RADUX equipment is synchronized and 15 minutes is allowed for the phase indicator to arrive at a stable reading.

Switch S2, which supplies 10 microvolts of white noise in a 600-cps bandwidth to the RADUX receiver input, is closed and the reading of line-of-position No. 1 is recorded at 1/2-minute intervals for 30 minutes. The average reading for the 30-minute period is calculated by adding all the readings and dividing by the total number of readings.

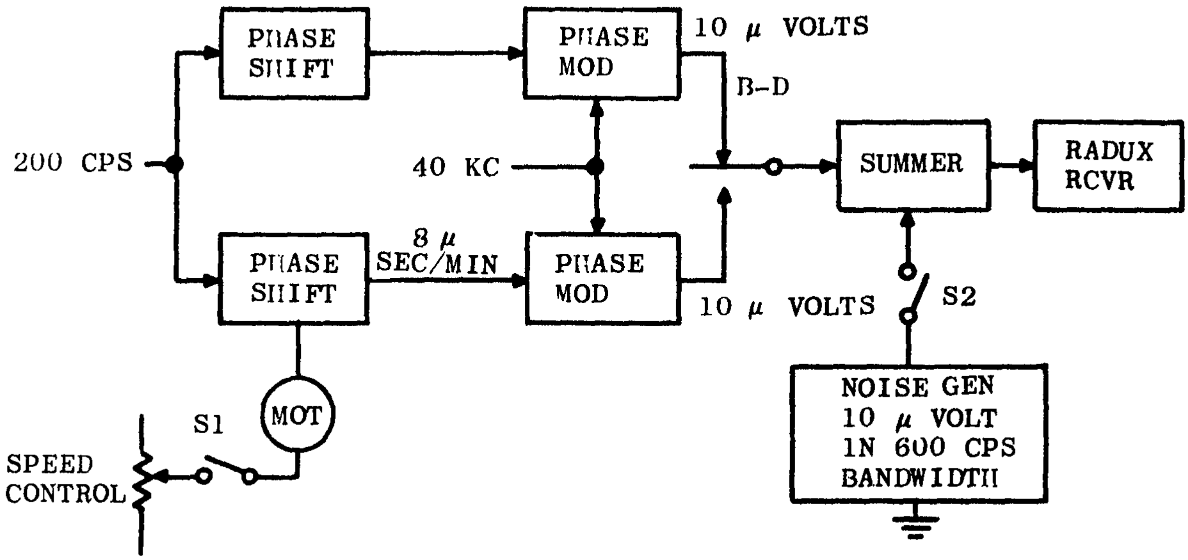


Figure 5. Tracking Test, Block Diagram

Switch S1, which changes the phase of one of the signals at an 8-microsecond-per-minute rate, is then closed and the reading of line-of-position No. 1 is recorded at 1/2-minute intervals for 30 minutes. The lag is calculated by using the following formula:

$$\text{Lag} = \sum_{i=1}^{n} \frac{(X_i - \bar{X}) - \text{input}}{n} \qquad (3)$$

where:

X_i = any single reading,

$\bar{X}$ = average reading, and

n = total No. of readings.

2.4.6 Sensitivity (fig. 6)

Switch S1 is placed in the position that connects the 500-ohm line to the signal generator. The signal generator output is adjusted to zero and the voltage at M2 is measured with the loop switch in each of its three positions, OMNI, FORE-AFT and ATHWART. The signal generator output is then adjusted until the voltage at M2 is ten times the level measured with zero signal input. The loop is oriented for maximum output for each position of the loop switch. The signal generator output for each loop switch position is the sensitivity under those conditions.

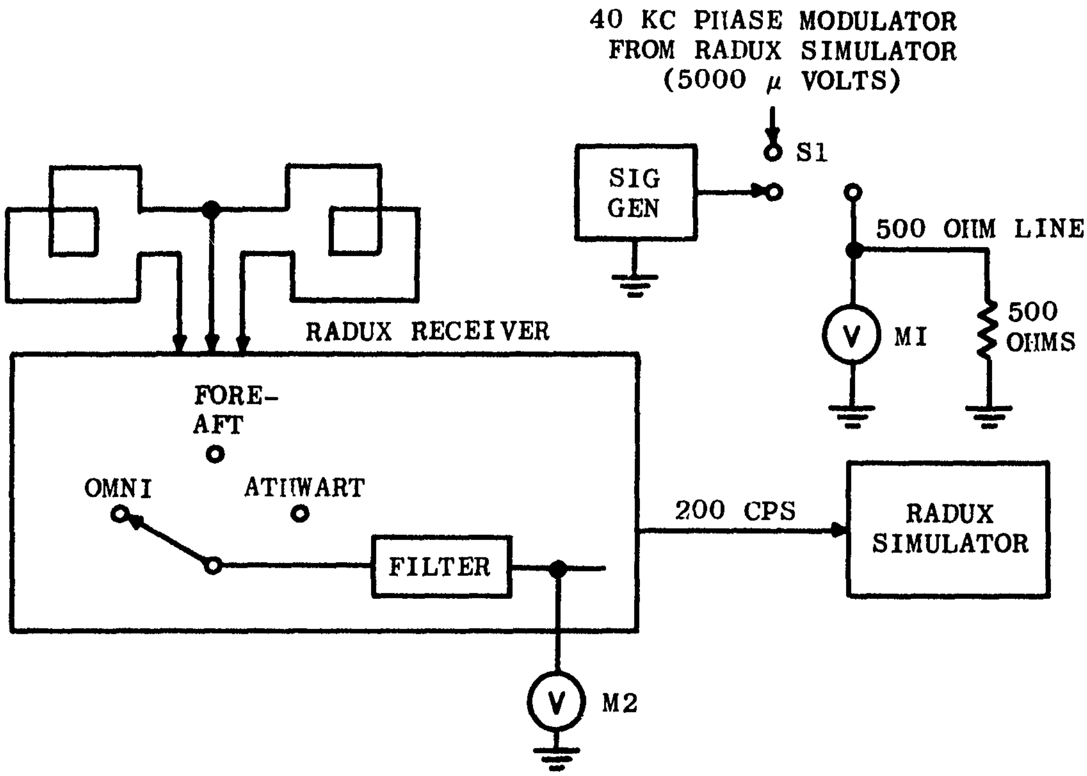


Figure 6. Sensitivity Test, Block Diagram

2.4.7 Omni-Test (fig. 7)

Switch S1 is placed in the position connecting the 500-ohm line to the RADUX signal simulator. The RADUX equipment is synchronized and 15 minutes are allowed for the reading to stabilize with the loop switch in OMNI position. The differential is disconnected from one of the gear trains and fastened on the left side so that it cannot move. The loop is rotated 360 degrees in 45-degree steps and the line-of-position reading at each position is recorded. Five minutes are allowed for stabilization at each position.

2.4.8 Dynamic Range and Maximum Signal Input (fig. 8)

The simulator supplies a 200-microvolt signal during the B and D transmission periods and a selectable amplitude signal with a 100-db range during the A and C periods. The RADUX equipment is synchronized and line-of-position No. 1 is read

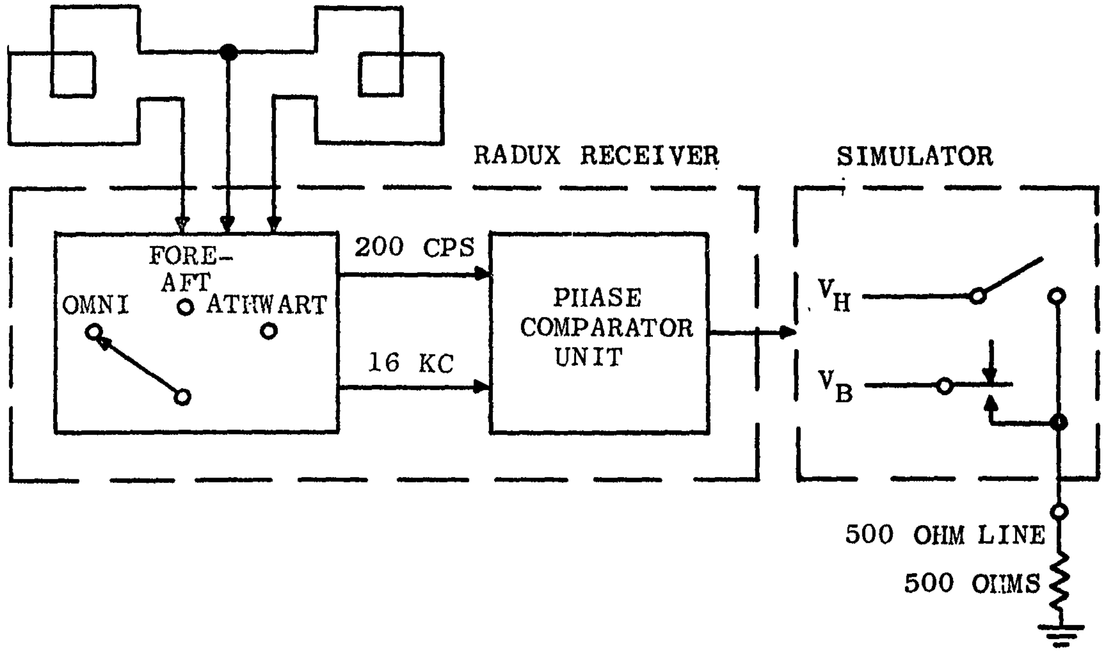


Figure 7. Omni Test, Block Diagram

for each signal-level input, which is varied from 3 to 300,000 microvolts in 10-db steps. Five minutes are allowed for each reading following the initial reading.

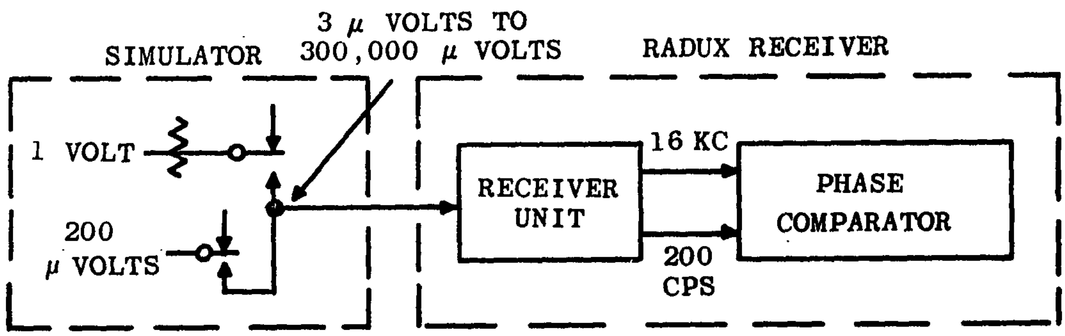


Figure 8. Dynamic Range And Maximum Signal Input Test, Block Diagram

3. DETAIL FACTUAL DATA

3.1 RECEIVER UNIT

The basic receiver design was based on the three principles of good receiver design, i.e., (1) separation of the gain and selectivity circuits, (2) placement of the selectivity-determining circuits as far ahead of the gain circuits as possible (as near as possible to the antenna without degrading the noise figure), and (3) utilization of filters for determining selectivity rather than the conventional interstage-tuned circuits.

3.1.1 Gain Changing Circuits

One of the more severe requirements placed on the receiver is that it be capable of operating over a dynamic signal range of 100 db, i.e., good phase stability is required between two signals that can vary in amplitude by as much as 100 db during a single measuring interval (3 seconds). To ensure this operation, near-perfect signal limiters must be inserted in the receiver. Several techniques for phase-stable limiting were evaluated during the program.

3.1.1.1 Cathode-Coupled Limiter

The cathode-coupled limiter requires two triodes per stage. An input cathode follower is cathode-coupled to a grounded grid amplifier whereupon limiting is affected by over-driving the stage into its cut-off regions. By properly adjusting circuit parameters, symmetrical clipping may be accomplished. When the signal is too weak for limiting, the stage acts as an amplifier with typical measured gain values of from 12 to 16 db. In terms of balanced limiting and minimum phase shift, this circuit was adequate; but because of the low value of gain it required too many stages.

3.1.1.2 Switched Gain Control

The stage gain of pentode amplifiers was adjusted by changing the bias with a fixed potentiometer. Phase-shift-versus-amplitude tests were made with constant modulation-index f-m signals. Results of these tests indicated that differential phase shifts over the 100-db dynamic range can be kept below 10 microseconds. However, achievement of this low value of phase shift is dependent upon a very critical adjustment of the potentiometer, which along with the problem of commutating a different gain control for each incoming signal, made this solution undesireable.

3.1.1.3 Forward Gain Control

A method of controlling the gain in the foreward direction was evaluated. In this method, the signal at the plate of each stage is rectified and fed as a negative bias voltage to the succeeding

stage. Efficiency of gain control is improved by feeding only a portion of the signal voltage to the grid of the succeeding stage. The test receiver, with 6 stages, had a gain of 120 db. With a 100-db change in signal level, the output varied only 1 db and the change in time delay was only 9 microseconds. However, this method was abandoned because the recovery time between signals was more than could be tolerated and the fact that this technique is an open loop one and will require readjustments with different environments.

3.1.1.4 Diode Limiters

The best circuit tested, from both the phase-shift and economy-of-stages standpoints, utilized diode limiters and pentode amplifiers. A pair of high-conductance silicon diodes, placed back-to-back to limit both positive and negative signals, were used after each pentode stage. When saturated, the diodes limit the signal to approximately 1.5 volts peak-to-peak. This ensures that the amplifiers will always operate within their linear region. The basic interstage circuit is shown in Figure 9.

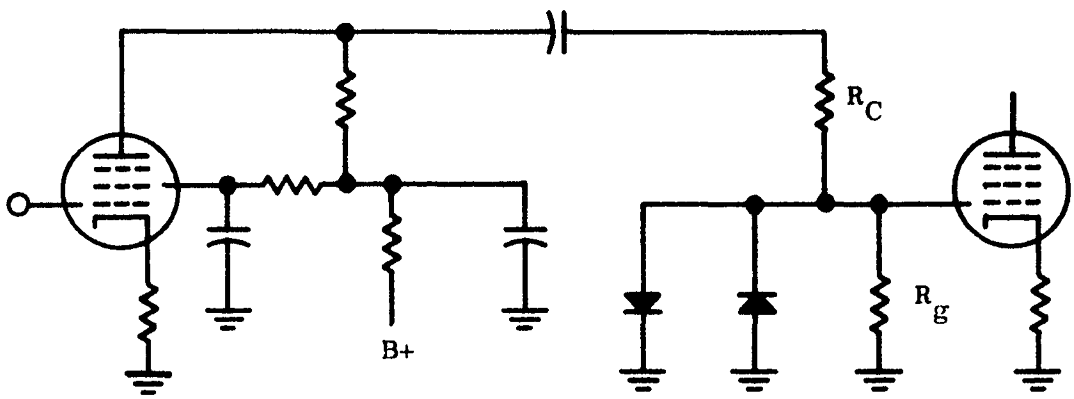


Figure 9. Interstage Limiting Circuit, Schematic

Each diode may be represented by a variable capacitor in parallel with a variable resistor. The variation of the capacity (C) across the diodes is a function of signal amplitude. R_g should be picked so that it is smaller than the impedance of the shunt capacity (X_c) to eliminate any differential phase shifts with amplitude. R_c isolates the diodes from the amplifier and should be very small compared to the impedance of the diode capacity.

Phase-shift-versus-amplitude tests using this method indicate maximum modulation phase shifts of 7 microseconds over a 100-db amplitude variation.

3.1.2 TRF vs Superheterodyne

A comparative analysis was made to determine whether TRF or superheterodyne techniques would be better for this development. The TRF technique offered some advantage in design simplicity since a mixer stage would not be required, but this advantage was overshadowed by the fact that the TRF would require all gain to be at one frequency and it would be extremely difficult to minimize unwanted feedback. Furthermore, even though both techniques require four different coherent signals, the TRF requires that the discriminator be retuned when the carrier frequency is changed. As a result of the analysis, it was decided that the superheterodyne technique would be best.

3.1.3 Loop Antenna Coupler

It was determined early in the program that the sensitivity and phase shift requirements of the loop input were very stringent and would require an extensive development program to achieve adequate results. Thus early development was centered around the use of an untuned transformer and a low-noise amplifier.

3.1.3.1 Grounded-Grid Amplifier

In an attempt to achieve a minimum noise figure, a high-transconductance, type 5842 triode was used as a grounded-grid amplifier with an untuned transformer input. Although fairly low equivalent noise resistances were achieved for the tube alone, the combined circuit proved ineffective. Overall sensitivity was poor due to the voltage loss that resulted from matching the relatively high antenna impedance to the low impedance amplifier input.

3.1.3.2 Cascode Amplifier

A cascode amplifier, that featured low equivalent noise resistance and gave good results with the untuned transformer, was designed. The basic amplifier is a grounded-cathode-grounded-grid connection. The grounded-grid stage drastically reduces capacitive feedback from output to input without introducing partition noise as would be produced in the screen current of a pentode. Shot noise, contributed by the grounded-grid stage, is negligible because of the plate resistance degenerative effect of the grounded-grid stage, that is in series with the cathode of the grounded-grid stage. The noise figure thus approaches the theoretical noise of the input section.

This circuit is valuable as a feedback amplifier because of the 180-degree relationship between input and output, and also because of the high gain. Values of gain between 46 and 54 db were measured for different tubes.

One disadvantage of this arrangement, however, is the large amount

of hum produced as a result of the cathode-to-filament rectification of the grounded-grid section. This effect is caused by the high positive voltage on the cathode.

The equivalent noise resistance (R_{eq}) of a triode is approximated by:

$$R_{eq} = \frac{2.5}{gm} \tag{4}$$

where:

gm is the operating transconductance.[1]

The measurement of equivalent noise resistance was accomplished as follows: the amplifier input was grounded and the amplifier output voltage, of a known bandwidth, was measured. The ground was then removed from the amplifier input, and a signal of such amplitude as to produce an amplifier output ten times as great as that obtained with the input grounded was then supplied. The equivalent noise resistance is found from the equation:

$$R_{eq} = \frac{E_n^{\ 2}}{4KTB} \tag{5}$$

where:

E_n = rms noise voltage,

K = Boltzmann's constant = 1.38×10^{-23} joules/degree,

T = temperature in degrees Kelvin and

B = noise bandwidth in cycles per second.

Using this method of measurement, the type 6922 cascode amplifier was found to have 230 ohms equivalent noise resistance, with an operating transconductance of 12,000 micromhos.

3.1.3.3 Untuned Transformer

A complete analysis was made of the untuned input transformer with assumptions pertaining to circuit parameters.[2,3] It was

---

[1] Reference Data For Radio Engineers (New York, 1956) pp

[2] Report On Noise In FM Systems And Loop Sensitivity In RADUX System (October, 1955), (Confidential)

[3] Interim Engineering Report No. 18 For Navigation Receiver For Use In RADUX LF Navigation System (March, 1959). (Confidential)

determined that the best that could be hoped for in terms of the specification sensitivity was 240 microvolts compared to the required 150 microvolts. Since the specification could not be met with the untuned transformer, other approaches were investigated.

3.1.3.4 Feedback Amplifier[4]

Regardless of the antenna coupler input circuit used, a highly stable amplifier is required between the antenna and the r-c coupler phase shifter. It has been shown that the carrier phase shift at the output of the r-c network is independent of loop orientation only if the gain and phase shift of each loop circuit is equal.[5] Since feedback amplifiers are characterized by an increase in input impedance, a decrease in output impedance and a decrease in gain gain changes due to power supply variations and circuit parameters, these properties are used to advantage in the loop antenna coupler. (Refer to Figure 10.) The effect of feedback, then, is to reduce the error caused by gain differences and to lower the output impedance of the cathode follower that is driving the r-c coupler circuit.

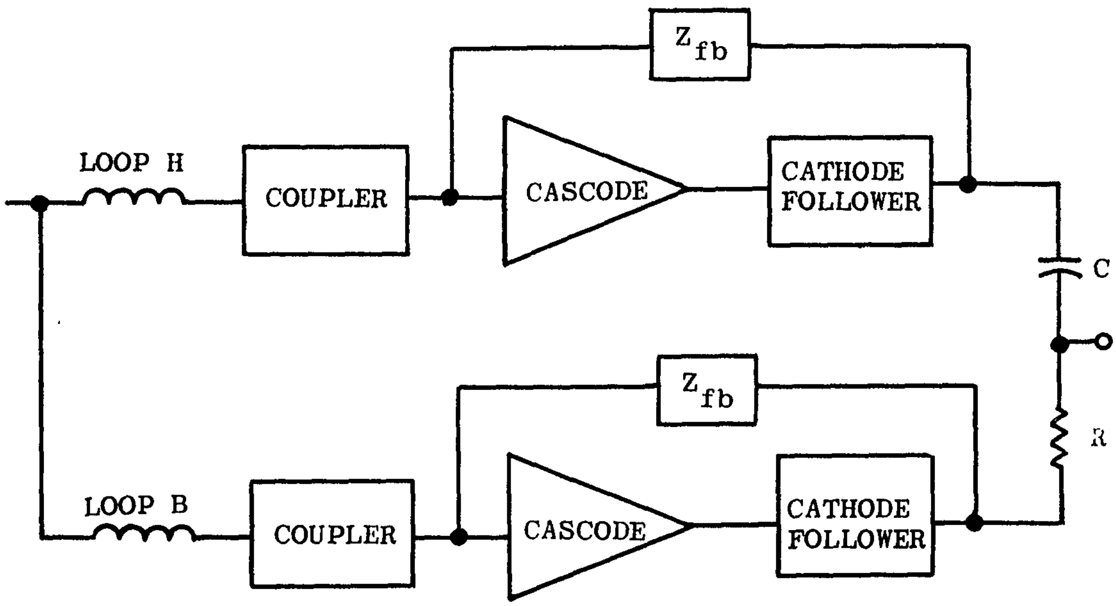


Figure 10. Basic Antenna Loop Coupler, Block Diagram

---

[4] Truman S. Gray, Applied Electronics (New York, 1954), pp

[5] R. J. Ertman, Investigation Of Underwater Radio Reception (Washington D. C., 1954), pp.

3.1.3.5 Tuned Loop

Maximum sensitivity may be obtained by stepping up the impedance of the loop to minimize the effect of the noise resistance of the first amplifier. A method for achieving this result without loss is obtained by tuning the loop to the carrier frequency.

Measurements made on a Navy AT-317/BRR Crossed Loop indicate loop inductance of 500 microhenries and a maximum Q of 115. The effective noise resistance (r) of the antenna is calculated to be:

$$r = \frac{\omega L}{Q} = \frac{(2\pi)(4\times10^{4})(5\times10^{-4})}{115} \approx 1.1 \text{ ohms} \qquad (6)$$

The noise (E_n) produced in the 600-cycle bandwidth by 1.1 ohms is:

$$\begin{aligned} E_n &= \sqrt{4KTBr} \\ &= \sqrt{(4)(1.38\times10^{-23})(3\times10^{2})(6\times10^{2})(1.1)} \qquad (7) \\ &= 3.3 \times 10^{-9} \text{ volts} \end{aligned}$$

For a signal-to-noise ratio of 10, the required signal voltage (E_a) induced in the loop equals 3.3×10^{-8} volts.

The effective height of the loop was found to be 2.2 millimeters. Sensitivity (E_L) of a 10-to-1 signal-to-noise ratio, as referred to the line, is calculated to be:

$$\begin{aligned} E_L &= \frac{(E_a)(7.75)}{(2.2)(10^{-3})} \\ &= \frac{(3.3)(10^{-8})(7.75)}{(2.2)(10^{-3})} \qquad (8) \\ &\approx 117 \text{ microvolts} \end{aligned}$$

Tuning increases the noise voltage by Q and its noise resistance by Q^2. Thus, using the measured value of Q and r, the noise resistance offered to the input stage is:

$$Q^2 r = (115)^2(1.1) \approx 14.6 \text{ K ohms} \qquad (9)$$

Sensitivity is degraded less than 1 percent by connecting this circuit to an amplifier with an equivalent noise resistance of 250 ohms.

Tuning both loops and connecting them for an omnidirectional pattern, the phase shift versus rotation was found to be on the order of 100 microseconds. Theoretically, this error could be made quite small if the gain and phase shift of each loop were equal. However, it would be an almost impossible task to make the the Q's of the respective loops and the tuning exactly equal and to maintain them equal in a varying environment.

3.1.3.6 Operational Amplifier

The advantages of the tuned loop can be utilized with few of the disadvantages by employing an operational amplifier as an input stage. Basically, the operation would be as illustrated in Figure 11 where E_a is the voltage induced into the antenna.

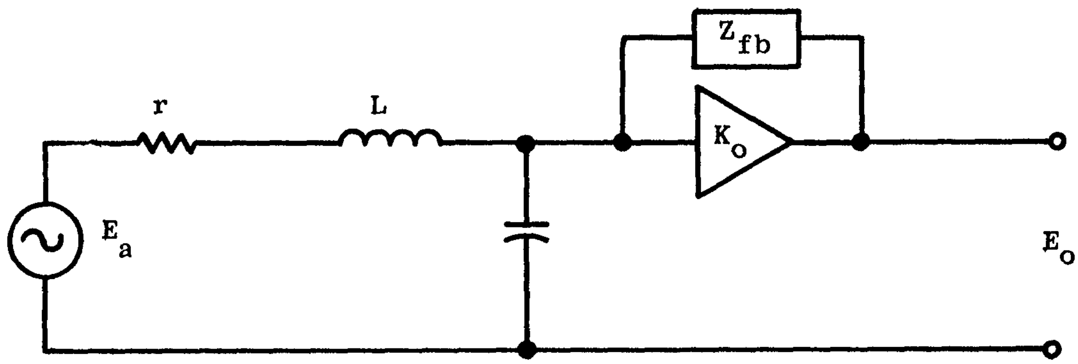


Figure 11. Operational Amplifier Input Stage

This may be reduced by Thevinin's equivalent as shown in Figure 12.

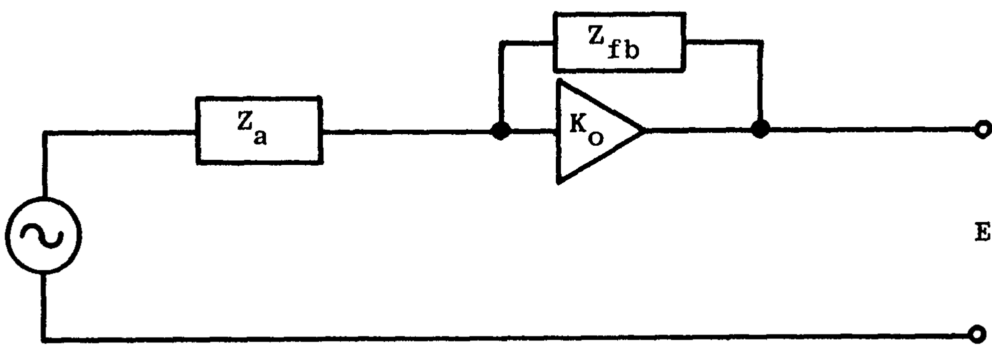


Figure 12. Equivalent Input Stage

$$\frac{E_o}{E_b} \approx \frac{-Z_{fb}}{Z_a} \qquad (10)$$

when:

$$K_o \gg \frac{Z_{fb}}{Z_a}.$$

$$Z_a = \frac{\frac{1}{j\omega c}(r+j\omega L)}{\frac{1}{j\omega c}+r+j\omega L},$$

which reduces to:

$$Z_a = \frac{-jrQ_o(1+jQ_o\frac{\omega}{\omega_o})\ \frac{\omega o}{\omega}}{1+jQ_o(\frac{\omega}{\omega_o} - \frac{\omega_o}{\omega})}$$

where:

$$Q_o = \frac{\omega_o L}{r} = \frac{1}{\omega_o cr},$$

ω_o = resonant frequency and

$$E_b = Ea\left[\frac{\frac{1}{j\omega c}}{(\frac{1}{j\omega c})+r+j\omega L}\right].$$

E_b then reduces to

$$E_b = Ea\left[\frac{-jQ_o(\frac{\omega_o}{\omega})}{1+jQ_o(\frac{\omega}{\omega_o} - \frac{\omega_o}{\omega})}\right].$$

Substituting for E_b:

$$\frac{E_o}{E_a} = \left[\frac{-jQ_o(\frac{\omega_o}{\omega})}{1+jQ_o(\frac{\omega}{\omega_o} - \frac{\omega_o}{\omega})}\right] \left[\frac{-Z_{fb}-1+jQ_o(\frac{\omega}{\omega_o} - \frac{\omega_o}{\omega})}{-jrQ_o(1+jQ_o\frac{\omega}{\omega_o})(\frac{\omega_o}{\omega})}\right],$$

which reduces to

$$\frac{E_o}{E_a} = \frac{-Z_{fb}}{r(1+jQ_o\ \frac{\omega}{\omega_o})}.$$

The phase angle $\emptyset$ is given by

$$\emptyset = \pi - \tan^{-1} Q_o\ (\frac{f}{f_o}) \tag{11}$$

when:

$$Z_{fb} = R_{fb}$$

and the phase slope

$$\frac{d\emptyset}{d_f} = \frac{\frac{-Q_o}{f_o}}{1+Q_o{}^2(\frac{f^2}{f_o{}^2})} \tag{12}$$

The maximum phase slope occurs when $f = f_o$, i.e.,

$$\left.\frac{d\emptyset}{df}\right|_{f=f_o} = \frac{\frac{-Q_o}{f_o}}{1+Q_o{}^2} \approx - \frac{1}{Q_o f_o} \text{ radians/cycle} \tag{13}$$

By properly choosing Z_{fb}, the phase slope can be made zero. However, a resistor was found to work well in this application.

By a similar analysis, the phase slope of the tuned loop without feedback is found to be

$$\left.\frac{d\phi}{df}\right|_{f=f_o} = - \frac{2Q_o}{f_o} \text{ radians/cycle} \qquad (14)$$

Since the differential time delay is directly proportional to the phase slope, a trememdous improvement is achieved by using feedback. Test results using this method yielded a sensitivity of 135 microvolts in planar and 190 microvolts in omni and a differential phase delay of 3 microseconds for 360-degree rotation.

3.1.4 Whip Antenna Coupler

In the analysis that follows, the antenna is assumed to consist of a generator of voltage E_a with an internal impedance Z_a[6]. At 40 kc, this impedance is almost all capacitive; therefore, the assumption is made that Z_a equals $1/j\omega c_a$.

Because of the 150-foot separation between the antenna and the receiver, a double-tuned circuit, with the primary at the antenna, is suitable in this application. Two possible configurations considered were (1) capacitance coupling and (2) mutual-inductance coupling.

In the capacity-coupled circuit (fig. 13) the capacity, C_m, would be made up partly, or wholly, of the connecting cable capacity. Therefore, its coupling would be a function of cable length requiring a separate adjustment for C_m.

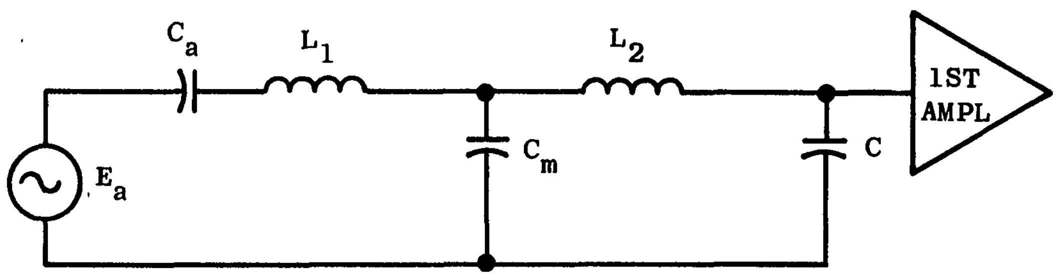


Figure 13. Capacitor-Coupled Circuit, Schematic

In the mutual-inductance (transformer) coupled circuit (fig. 14), the coefficient of coupling may be fixed, independent of cable length.

---

[6] K. R. Sturley, Radio Receiver Design (New York, 1953) pp.

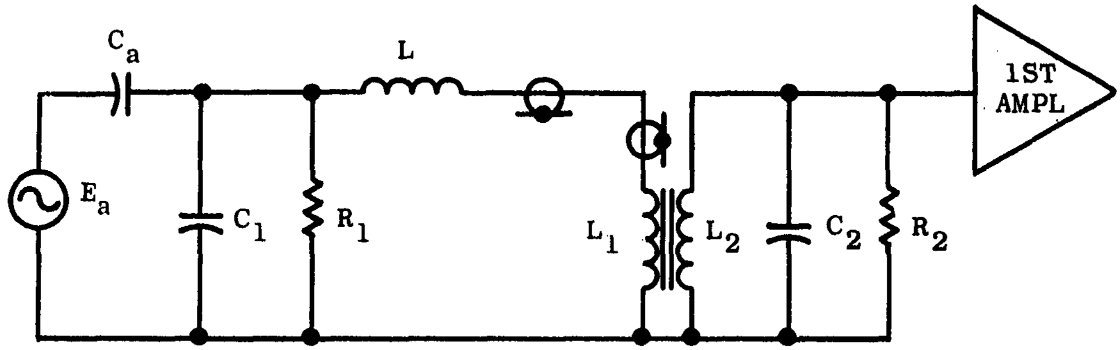


Figure 14. Transformer-Coupled Circuit, Schematic

If the assumption is made that the transformer has a coefficient of coupling (k) equal to 1, the effective circuit coefficient of coupling is:

$$k_{eff} = \sqrt{\frac{L_1}{L+L_1}} . \tag{15}$$

For maximum flat response at the output,

$$Q\ K_{eff} = 1 \tag{16}$$

The bandwidth of the circuit was picked to be approximately 4 kc. This makes a comparatively wide range of flat phase slope. Small change in time delay is expected for changes in whip capacity and circuit parameters.

For the transformer-coupled application, L was chosen to be 10 millihenries, L_1 to be 80 microhenries and L_2 to be 80 millihenries. C_1 is large compared to C_a and small variations in C_a will not have much effect on the time delay.

$$\frac{E_2}{E_a} = \frac{Z_T}{Z_a} \tag{17}$$

where:

Z_T = transfer impedance

If Q_P equals Q_S equals Q equals 12 and the circuit is designed to make Qk equal to 1, then the transfer impedance is given by

$$Z_T \text{ max} = \frac{\sqrt{R_{DP} R_{DS}}}{2} \tag{18}$$

where:

R_{DP} = dynamic impedance of primary circuit,

R_{DS} = dynamic impedance of secondary circuit and

$$Z_{T\ 40\ kc} = \frac{\sqrt{(31 \times 10^3)(2.4 \times 10^5)}}{2} = 4.3 \times 10^4 \text{ ohms}$$

The noise produced by this circuit in a 600-cps bandwidth, neglecting the 1st r-f amplifier noise, is

$$\begin{aligned} E_n &= \sqrt{4KTBR} \\ &= \sqrt{(4)(1.38\text{x}10^{-23})(3\text{x}10^2)(6\text{x}10^2)(4.3\text{x}10^4)} \\ &\approx .65 \text{ microvolts.} \end{aligned} \tag{19}$$

The signal (E_2) required for a signal-to-noise ratio of 10 is

$$E_2 = 6.5 \text{ microvolts.} \tag{20}$$

The voltage (E_a) induced at the antenna for sensitivity:

$$\begin{aligned} E_a &= \frac{E_2 Z_a}{Z_T} \\ &= \frac{(6)(2\text{x}10^4)}{4.3\text{x}10^4} \\ &= 2.8 \text{ microvolts} \end{aligned} \tag{21}$$

By letting Q_P approach infinity and changing Q_S for transitional coupling, the sensitivity may be improved by 3 db. In this case

$$E_a = (2.8\text{x}10^{-6})(.707) = 2 \text{ microvolts} \tag{22}$$

Comparative phase measurements indicate approximately 2 microseconds phase shift attributed to the whip coupler over the 100-db input range.

3.1.5 Mixer

A type 6BE6/5750 pentagrid is used as a mixer to change the rf to the 16-kc if. The circuit components were optimized for best amplitude versus-phase characteristics and conversion gain. The best results obtained were with approximately 1.6 volts local-oscillator drive and 2 volts bias. Under these operating conditions, the conversion gain is approximately 12 db.

3.1.6 Filters

3.1.6.1 R-F Filter

The r-f filter is a bandpass filter with the following specifications.

(1) Center frequency capable of being switched to 40 and 34 kc.

(2) Minus 3-db bandwidth of 600 ±60 cps.

(3) Minus 60-db bandwidth of not greater than 4000 cycles.

(4) Input impedance (external to the filter) of 10 k.

(5) Output impedance (internal to the filter) of 10 k.

(6) Insertion loss of between 10 and 14 db.

(7) Shift in center frequency of not more than ±20 cps over an operating temperature range of 0 to 70 degrees C.

A major problem encountered in designing this filter was the correction of the time delay differential at the sideband frequencies with changes in input amplitude. This time delay was considered to be affected primarily by the inductance changes occurring because of non-linear B-H characteristics of the iron cores used in the filter. Consequently, the cores were made as large as feasible with respect to size and weight.

The differential time delay, with respect to the 100-db input amplitude change at the f_o ±200-cps frequency points, was calculated to be no greater than 2 microseconds. Measurements in the system indicate that no more than this amount could be attributed to the filter.

3.1.6.2 I-F Filter

The i-f filter is a bandpass filter with the following specifications:

(1) Center frequency of 16 kc.

(2) Minus 3-db bandwidth of ±750 ±10 cps.

(3) Minus 60-db bandwidth of not greater than ±3000 cps.

(4) No flyback above minus 60 db.

(5) Input impedance (external to the filter) of 18 k.

(6) Output impedance (internal to the filter) of 18 k.

(7) Insertion loss no greater than 5 db.

(8) Shift in center frequency of not more than ±20 cps over an operating temperature range of 0 to 70 degrees C.

The differential time delay, with respect to the 100-db input amplitude change at the f_o ±200-cps frequency points, was calculated to be no more than 0.5 microseconds.

3.1.6.3 Local Oscillator Filter

The local oscillator filter is a bandpass filter with the following specifications:

(1) Center frequency of 52.5 kc.

(2) Three-db bandwidth of approximately 8 kc.

(3) At least 40 db rejection for frequencies more than ±7 kc away from the center frequency.

(4) Input impedance of 10 k.

(5) Output impedance of 50 k.

The primary function of the filter is to reject unwanted signals that may be produced in the frequency divider and to eliminate any 16 kc that may be picked up within the receiver.

3.1.6.4 Low-Pass Filter

The low-pass filter has the following specifications:

(1) Minus 3 db at 500 cps.

(2) Input and output impedances of 10 k.

(3) At least 40 db rejection for frequencies more than 200 cps above the cutoff frequency.

This filter is located at the output of the discriminator and has a bandwidth that is wide enough to pass the effective receiver noise bandwidth. A cutoff frequency of 500 cps was considered to be wide enough.

3.1.6.5 Crystal Filter For Synchronizer

The crystal filter for the synchronizer has the following specifications:

(1) Center frequency of 16 kc ±1 cps.

(2) Six-db bandwidth of 12 cps.

(3) Sixty-db bandwidth of 40 cps.

(4) Input and output impedances of 10 k.

(5) Insertion loss of less than 5 db.

(6) Shift in center frequency of not more than ±1 cps over an operating temperature range of from 0 to 60 degrees C.

The bandwidth was made as narrow as the associated rise time would allow. The rise time causes a shift in the signal pattern; a shift that cannot be so great as to cause the information channel to sample at the wrong time.

3.1.7 Discriminator

At the beginning of the RADUX program, it was decided that coherent detection did not offer enough improvement over conventional Foster-Seeley type discriminators. This conclusion was based on NRL report no. 435 and conferences with NEL and NRL personnel.

In the analysis of this system, it was found that the discriminator degraded the signal-to-noise ratio. In the worst case, i.e., with a signal-to-noise ratio of 1 to 2, the output of the discriminator was measured with an approximate signal-to-noise ratio of 1 to 4. The phase detector loop bandwidth was determined from the tracking rate required and the worst signal-to-noise ratio to be encountered. The calculations were based on an input signal-to-noise ratio to the phase detector of 1 to 4 and a maximum phase error of 10 microseconds. Improving the signal-to-noise ratio into the detector, it was found, would not improve the phase accuracy much as long as the phase detector was operating in a linear region.

The only advantage to coherent detection, therefore, would be the improvement of the signal-to-noise ratio into the phase detector, which would result in a smaller standard deviation. However, this type of operation would require separate phase loops, one for each transmitting station, and so was not considered practical for this model.

Two conventional-type discriminators were tried; the Travis type and the Foster-Seeley type. It is possible to show that, electrically, these circuits are substantially the same.[7]

The Foster-Seeley discriminator requires a double-tuned transformer with a fixed coefficient of coupling. A property of

---

[7] L. B. Argiumbau, Vacuum Tube Circuits (New York, 1948) pp.

double-tuned transformers is that the voltage across the tuned secondary will have a phase angle of 90 degrees relative to the voltage across the tuned primary and that the phase of the secondary voltage will swing up to plus and minus 90 degrees from this right-angle relationship as the frequency deviates below or above the center frequency. This action changes the fm to an amplitude variation and is recovered by a peak-detecting circuit resulting in the familiar "S" curve.

The Travis discriminator uses two tuned circuits, one tuned above the center frequency and one tuned below it. The action of this circuit is to convert the fm to am and then detect the modulation by peak-detecting diodes. The difference between these two outputs results in the discriminator "S" curve.

The linear region of the "S" curve should be wide enough to pass the receiver noise bandwidth since a narrow linear region could cause discriminator unbalance under noisy conditions.

Noise balance is critical under poor signal-to-noise ratios. Unbalance can cause excessive shifts in mean when the signal is reduced from good signal-to-noise ratios to poor ones. The major cause of noise unbalance is the receiver noise bandwidth not being symmetrical about the center frequency.

3.1.8 Receiver Power Supply

The receiver power supply is a 250-volt, 150-milliampere regulated supply. The output impedance is approximately 1 ohm with no more than 2 millivolts ripple under full load conditions. The 115 volts ac supplied is controlled by the power switch, which is located on the phase comparator unit.

The power transformer has two 6.3-vac taps for the filaments, a 3-amp tap used for the power supply tubes exclusively and a 6-amp tap for the receiver tubes.

3.2 PHASE COMPARATOR UNIT

3.2.1 Phase Detector

The basic phase detector is a half-wave, two-diode phase-measuring device. This type of detector was chosen because of its simplicity and ease of balancing. The average output voltage is zero when the received signal is 90 degrees out of phase with the reference signal and is either positive or negative when the received signal leads or lags the basic 90-degree phase difference. The average d-c output is extracted from the phase detector and developed across an integrating capacitor.

The integrator is a storage device with its signal applied to one contact of the chopper input to the servo amplifier. An oscilloscope is used to adjust the average level of the phase

detector for zero with no received-signal input.

An undesirable effect of using this technique is that grid current from the servo-amplifier input stage charges the integrator capacitor and yields an output error when no actual error exists. However, this effect has been eliminated by applying a voltage, which is equal to the undesired error voltage, to the opposite chopper contact.

3.2.2 Servo Amplifier

The servo amplifier is a conventional amplifier with a push-pull output driving a two-phase, servo-motor load. Feedback is employed to reduce the output impedance of the amplifier and to stabilize the gain. The closed-loop amplifier gain is approximately 2000. The input stage has been designed to minimize grid-current flow.

3.2.3 Phase Shifter

The phase shifter utilizes a precision resolver in conjunction with a resistor and capacitor. The resistor is adjusted so that the resistance is equal in magnitude to the capacitive reactance. Ideally, the output of the phase shifter is constant in amplitude and its phase is linearly related to the mechanical rotation of the rotor. The resolvers alone contribute a peak-to-peak error of 5 microseconds in respect to a perfectly linear relationship between output phase and rotor mechanical rotation.

Linearity tests show that the peak-to-peak error of the shifters together with the rest of the phase-measuring loops under static conditions (i.e., one signal at a time with no time sharing) is approximately 7 microseconds. The same tests made under dynamic (time sharing) conditions show the peak-to-peak error to be approximately 12 microseconds. The larger errors under dynamic conditions are due to cross-coupling between channels.

3.2.4 Fail-Safe Circuits

Fail-safe circuits are provided to indicate correct operation by illuminating the various lamps when the equipment is operating properly. The same basic design, consisting of a twin triode and a relay with an adjustment available to obtain circuit operation at the desired error signal amplitude, was used for all circuits.

3.2.5 Local Oscillator

Several local oscillator designs were evaluated and were found to be lacking in basic stability because of on-off heater controls being used for the crystal oven.

The Manson, precision, one-megacycle oscillator with proportional oven-temperature control was chosen because it guaranteed that the

long-time drift would not exceed 1 part in 10^8 per day. Short-time drift is minimized by the elimination of oven temperature cycling. Tests on this unit have confirmed long-term drift figures of 3 parts in 10^9 per day and peak-to-peak short-time drift figures of approximately 1.5 parts in 10^9.

An investigation of the results of oscillator drift upon the operation of the equipment shows that an oscillator with a stability of 1 part in 10^8 per day will allow continuous operation without automatic frequency control (afc) 10 times longer than a stability of 1 part in 10^6. Maximum operating time without using afc can be calculated by letting:

θ_L = linear range of phase measuring servomechanism = 1.33 degrees,

θ_e = phase error when tracking at 40 knots along a baseline = 0.33 degrees, and

θ_{max} = maximum allowable phase error.

Then:

$$\theta_{max} = \theta_L - \theta_e \tag{23}$$

$$= 1.33 \text{ degrees} - .33 \text{ degrees}$$

$$\theta_{max} = 1 \text{ degree.}$$

Drift is attributed to both long- and short-time components, i.e.,

$$\theta_{max} = \theta_{long\ time} + \theta_{short\ time}. \tag{24}$$

$$\theta_{long\ time} = \frac{mt^2}{2} \tag{25}$$

where:

m = drift rate in degrees per second per second, and

t = time. (26)

$$(\theta_{short\ time})_{max} = \frac{AT}{\pi}$$

where:

A = peak short-time frequency drift, and

T = period of short-time drift.

Assuming a drift rate of 1 part in 10^8 per day and a peak short-timeddrift of 3 parts in 10^9:

$$m = \left(\frac{3.6}{4.32}\right) 10^{-8} \text{ deg/sec}^2,$$

$$A = (21.6)(10^{-5}) \text{ deg/sec, and}$$

$$T = 1800 \text{ sec (measured)}.$$

Thus, referring to equation (26),

$$(\theta_{\text{short time}})_{\text{max}} = (21.6 \times 10^{-5} \text{ deg/sec})(1.8 \times 10^3 \text{ sec}) \div \pi$$

$$= (12.4)(10^{-2}) \text{ deg}$$

$$= 0.124 \text{ deg},$$

and to equation (24).

$$\theta_{\text{long time}} = \theta_{\text{max}} - (\theta_{\text{short time}})_{\text{max}}$$

$$= 1 - .124$$

$$= .876 \text{ deg}.$$

Furthermore, by equation (25):

$$(t_{\text{max}})^2 = 2\theta_{\text{long time}} \div m$$

$$= \frac{(2)(.876)(4.32)}{(3.6)(10^{-8})} \text{ sec}^2$$

$$= (2.1)(10^8) \text{ sec}^2,$$

Therefore:

$$t_{\text{max}} = (1.45)(10^4) \text{ sec}$$

$$= (2.42)(10^2) \text{ min}$$

$$= 4 \text{ hours}.$$

3.2.6 Frequency Divider

The stability, versatility, reliability, ease of adjustment, and maintainability of both sine-wave and the binary feedback-type dividers were considered.

The binary divider is generally superior to the sine-wave type as far as stability and ease of adjustment are concerned. In both

types, however, it is relatively difficult to isolate troubles.

The sine-wave divider is extremely versatile in that the output frequencies are relatively easy to change. The fact that the frequencies are sine waves makes them easier to handle. Stability is adequate for difference-in-phase measurements and reliability is very good.

Thus, the sine-wave divider was chosen for its versatility and adequate stability characteristics, which make it adaptable, with only minor changes, to either a TRF or a superheterodyne receiver. Drift tests of the final sine-wave divider showed a maximum short-time drift rate of only 1 microsecond per minute while temperature tests on the prototype divider showed it was stable up to 80 degrees C.

3.2.7 Quadrature Coherent Detector

The quadrature coherent detector is used to extract the amplitude envelope of the transmission schedule from the received signal. This envelope is then used to synchronize the local commutator.

The detector itself is actually four separate phase detectors with four reference signals phased in quadrature. The output is the sum of the outputs of the four phase detectors. The purpose of the quadrature reference signals is to produce a relatively constant output regardless of the phase of the signal being detected.

Figure 15 shows how the amplitude of the quadrature coherent detector varies as the phase of the input signal is varied through

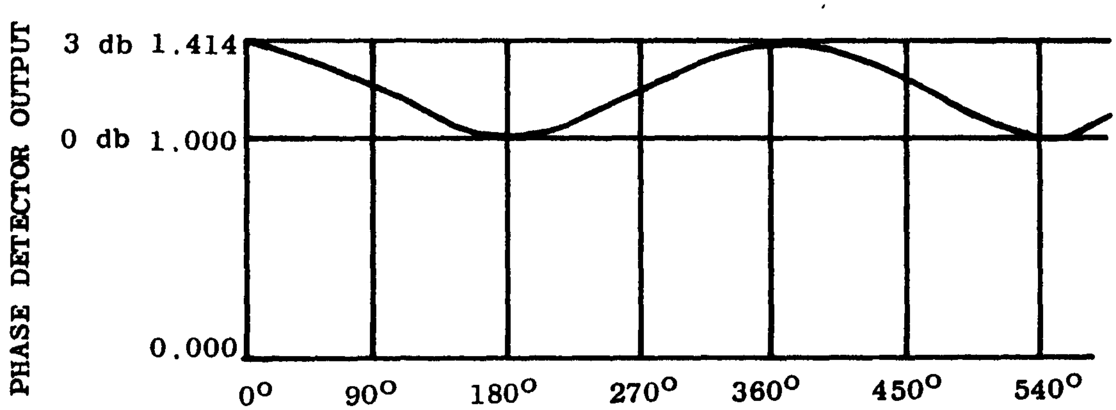


Figure 15. Quadrature Coherent Detector Amplifier vs Phase

360 degrees. The output amplitude of the detector can vary 3 db with the difference in frequency between the reference and the incoming signal determining the rate of change of the variation. To avoid this effect, the reference frequency was changed to approximately 20 cps lower than the incoming frequency. The frequency of the amplitude variations was then about 20 cps, which by using a relatively simple low-pass filter, was filtered out leaving a constant amplitude signal.

Detection efficiency tests with respect to signal-to-noise ratio were made between the quadrature detector and a plate detector and the quadrature detector was found to be about 3 db better.

3.2.8 Commutation Synchronizer

The commutation synchronizer circuit is used to synchronize the locally-generated gate signals with the detected transmission schedule envelope. The detected envelope must have a larger amplitude when a signal is present than it has when noise alone is present. To provide a reasonable degree of amplitude difference, the required ratio between the detected "signal plus noise" and "noise alone" signals is two, i.e.,

$$\frac{2}{1} = \sqrt{\frac{S^2 + N^2}{N'}} \tag{27}$$

Then, solving for the signal-to-noise ratio (S/N),

$$S/N = 3/1.$$

Therefore, a 3-to-1 signal-to-noise ratio is required at the output of the detector.

With a modulation index of $\pi/4$, 8/10ths of the rms signal voltage is due to the carrier frequency. For a signal-to-noise ratio of 1/2 in a 600-cps bandwidth, the signal-to-noise ratio at the receiver output is 0.4. Thus, to obtain the required 3-to-1 signal-to-noise ratio at the detector output, the bandwidth must be narrowed as follows:

$$S/N_{out} = S/N_{in}\sqrt{\frac{BW\ in}{BW\ out}} \tag{28}$$

$$3/1 = 2/5\sqrt{\frac{600}{BW}}$$

$$\frac{1}{BW} = \left(\frac{15}{2}\right)^2\left(\frac{1}{600}\right) = \frac{225}{2400}$$

$$BW = \frac{2400}{225} = 10.5 \text{ cps}$$

If the signal is passed through a 10.5-cps filter, the S/N output should be 3 to 1, which is the minimum required for synchronizing. However, passing the received signals through a 10.5-cps bandwidth filter at the 16-kc if causes the envelope to have a rise time of approximately 100 milliseconds. This rise time distorts the incoming envelope and delays it by approximately 100 milliseconds causing synchronization to be off by that amount. But, this is not serious since the maximum delay that is tolerable for proper channeling of the signals to the phase-measuring circuits is 120 milliseconds.

Thus, the narrowest filter usable for synchronization is approximately 10 cps. At the same time, a 10-cps bandwidth has been shown to reduce a 1-to-2 signal-to-noise ratio in a 600-cps bandwidth phase-modulated input signal to 3 to 1 at the output. These results lead to the conclusion that the lowest signal-to-noise filtering is 1 to 2 in a 600-cps bandwidth.

3.2.8.1 Synchronizing Under Poor S/N Conditions

Even through the signal input to the commutation synchronizer is held at a relatively constant level, there is still a difference in signal level under poor signal-to-noise-ratio conditions. This difference is due to the fact that, under noisy conditions, the total level, which includes both signal and noise, is held constant and the signal part is at a lower level.

A scheme of using agc under poor signal-to-noise-ratio conditions was tested but failed to operate properly. This scheme was unusable for the following reasons:

(1) Response time of the agc would have to be less than 120 milliseconds to prevent the synchronization point from shifting more than 120 milliseconds.

(2) Since response time is less than 120 milliseconds, the noise gain will automatically be controlled between transmissions in such manner that will tend to make the average noise level as high as the signal level.

(3) Commutated agc is worthless unless synchronization has already been accomplished by other means.

Therefore, a constant-amplitude technique using positive and negative limiting as shown in Figure 16 was designed.

This system is acceptable unless all stations have a 1-to-1 signal-to-noise ratio, at which time the sensitivity is not quite adequate. A good/poor switch has been incorporated to raise the signal level during those times when the signal-to-noise ratio is 1-to-1 or worse at all stations.

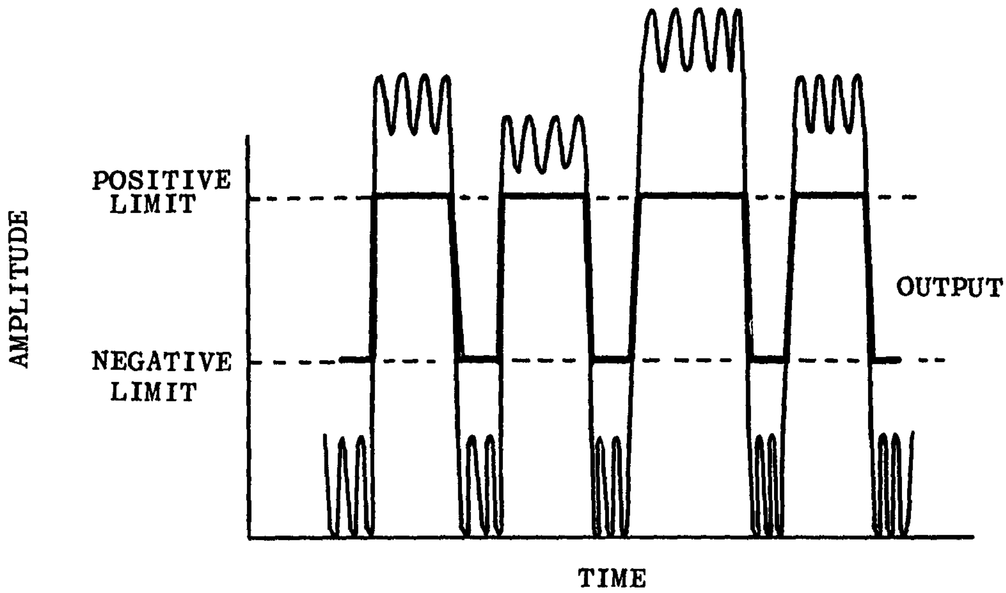


Figure 16. Constant-Amplitude Technique

3.2.8.2 Synchronizer Time Constant

To accomplish synchronization in 3 minutes, only four signal samples are allowed to bring the synchronizer up to the threshold level. This means that a relatively short time constant must be used and under poor signal-to-noise-ratio conditions, noise bursts can cause the threshold level to vary. After synchronization has been accomplished, these noise bursts can drive the system out of synchronization. Thus, to prevent this effect a longer time constant is switched into use when synchronization has been accomplished.

3.2.9 Electronic Commutator

The electronic commutator supplies all the gate signals that are used in the RADUX equipment. The timing reference for this unit is a 200-cps signal derived from the stable one-megacycle oscillator.

The commutator gates are used to operate relays that channel the received information to the proper phase-measuring loops. To ensure maximum reliability, extra-long-life relays are used. A life test on one of these relays operating at nine times the normal rate, was started in December, 1957. To date it has undergone 150 million operations and it is still operating satisfactorily.

3.2.10 Monitor Oscilloscope

The monitor oscilloscope is used to observe the receiver i-f

output. In this manner, the operator is able to determine whether or not the local oscillator is synchronized with the incoming signal. Since the sweep frequency is derived from the local oscillator, the pattern will remain stationary only when synchronization is accomplished.

A synchronizer lamp, which can be selected to light on any desired commutator segment, is used in conjunction with the oscilloscope to evaluate commutator synchronization.

3.3 POWER SUPPLY UNIT

The power supply unit for the phase comparator is contained in a separate case. It is a modified version of circuit No. 3A from the Handbook of Preferred Circuits prepared by the National Bureau of Standards. The supply will deliver about 1.5 amperes and the output variation is less than ±1 percent for input variations of ±10 percent.

3.4 REMOTE MONITOR UNIT

The remote monitor is a separate unit that can be used at a remote location to ascertain correct operation of the basic equipment. In addition to a monitor oscilloscope, identical to the one used in the phase comparator, it contains duplicates of all fail-safe lamps and a synchronizer-indicator lamp. The unit is a monitor only and cannot be used to control the main unit.

3.5 REMOTE INDICATOR UNIT

The remote indicator unit indicates the same basic line-of-position data as the phase comparator unit. It is small in size and can be used at a remote location to yield the basic position information. It contains synchro receivers coupled to veeder-root type counters for readout.

3.6 RECORDER UNIT

The navigation recorder is purchased from the Brush Instruments Division of Clevite Corporation. Line-of-position data is derived from synchro transmitters attached to the gear units in the phase comparator. The recorder has a synchro-receiver input with a servo follower to operate the recorder.

3.7 SEQUENTIAL CORRELATOR

The sequential correlator is a storage device using the theories that noise voltages add as the square root of the sum of the squares and that signal voltages add algebraically to increase signal-to-noise ratio.[8] Storage is accomplished by using a magnetic drum

---

[8] J. V. Harrington, "An Analysis of the Detection of Repeated Signals In Noise By Binary Integration", IRE Transactions On Information Theory, (March, 1955), pp.

rotating at a constant speed such that one revolution occurs for each two periods of the RADUX transmission sequence (one revolution per six seconds). Correlation is accomplished by adding succeeding signals to the stored information on the drum. However, the drum speed must be extremely constant and exact so that the signals will add to the stored information at the proper time. Correlation occurs when the drum speed is synchronized with the incoming signal to be stored. A 400-cps signal, derived from the local one-megacycle oscillator, is used to drive a synchronous motor and gear train coupled to the drum.

Measurements disclosed that drum speed was 2.0944 inches per second with variations of approximately $1.0472 \times 10^{-3} \sin 6.28t$ inches per second. As a result, the added signal is always within plus or minus 35 electrical degrees of the stored signal. Perfect correlation will not occur under these conditions, but the correlation is good enough to permit signal build-up on the storage medium.

Having established that the drum rotational velocity was adequate to produce correlation, the correlator was tested to determine whether the magnetic storage drum would add the signals as required. A precision, constant 200-cps signal was used to supply current to the recording head. This signal was derived from the same 1-mc oscillator as was the 400-cps signal used as the reference for the synchronous motor speed. A read-out head was used to determine the magnitude of the stored signal and it was observed that the signal did not increase with subsequent drum rotations. The record current was then doubled and the stored signal level was again observed and found to have doubled also. This test verified that the magnetic material was not saturated.

Other tests were made with different amplitudes, and at no time was a build up of stored information observed. The reason for this is that the magnetic field has the same strength during each application and the total field in existance on the tape cannot exceed the strength of the magnetizing force.

Another idea for accomplishing storage was advanced but lack of time has prevented its instrumentation. In this method, the tape would be used only as a delay device and the stored signal would be removed from the tape and added externally to the incoming signal. The combined signal would then be re-applied to the tape and stored again. In this manner, the signal could be built up over many measuring intervals.

3.8 ELECTROMECHANICAL DATA

3.8.1 Servo System

3.8.1.1 Bandwidth Considerations

The indicator bandwidth is specified in terms of the maximum lag

at 40 knots and the standard deviation with a 1-to-2 signal-to-noise ratio in a 600-cps receiver bandwidth.

The standard deviation (σ) is a function of the signal-to-noise ratio in the indicator bandwidth, i.e.,

$$\sigma = \tan^{-1} N/S \tag{29}$$

For very small angles (less than 3 degrees), the angle in radians is equal to the tangent of the angle,

$$\sigma \text{(in radians)} = \tan^{1} N/S = N/S \tag{30}$$

The specification requires a maximum standard deviation of 10 microseconds, which is 0.72 degrees. Therefore, the signal-to-noise ratio required in the indicator bandwidth is

$$\begin{aligned} S/N &= 1/\sigma \text{ (in radians)} \\ &= 180/.72\pi \\ &= 79 \end{aligned} \tag{31}$$

The discriminator causes approximately 10 db degradation of the signal-to-noise ratio. The maximum bandwidth of the indicator is then calculated as follows:

$$(BW_{servo})_{max} = \left\{ \frac{[S/N \text{ (in degradation)}]^2}{[S/N \text{ (required)}]^2} \right\} \left(\frac{BW_{input}}{2} \right) \tag{32}$$

$$= \left[\frac{(1/2)(1/3)}{79} \right]^2 (300) = .0013$$

In calculating the minimum bandwidth for tracking 40 knots, certain factors are considered. For instance, the tracking rate ($\dot{\theta}$) for tracking 40 knots is .068 microseconds per second. The integration gain (K_v) of the servo loop is then given by the equation[9]

$$\begin{aligned} K_v &= \dot{\theta}/\text{lag} \\ &= 0.68\ \mu\text{sec/sec}/10\ \mu\text{sec} \\ &= .0068/\text{sec} \end{aligned} \tag{33}$$

With a damping factor (ξ) of .707, the minimum bandwidth[10] to meet the tracking rate becomes

---

[9] Harold Chestnut and R. W. Mayer, <u>Servomechanisms And Regulating System Design</u> (New York, 1952), pp. 216.

[10] John G. Truxal, <u>Control System Synthesis</u> (New York, 1955)pp.293.

$$BW_{min} = K_v/\xi \tag{34}$$

$$= .0068/.707$$

$$= .0096 \text{ rad/sec}$$

$$= .00153 \text{ cycles/sec}$$

Since the signal input to the indicator system is sequentially time shared, the phase detector output is fed to the "A" channel integrator when the "A" station signal is present. The error information is stored during the time when the "A" station signal is not present. Each servo loop then gets its error signal from an integrator and the phase detector output is sampled sequentially once each 3 seconds by each integrator.

3.8.1.2 Servo Stability Analysis

For the RADUX system the time constant (T_1) is 1/2 and the open loop transfer function G(s), as a function of the complex frequency s, is given by the equation:

$$G(s) = \frac{K}{s(s + 1/2)} = \frac{2K}{s} - \frac{2K}{(s + 1/2)} \tag{35}$$

Where:

K = gain

Since the indicator is a sampled-data system, the analysis can be carried out in the Z plane as follows[11]:

$$G^*(s) = G(Z) = \frac{2KZ}{Z - 1} - \frac{2KZ}{Z - e^{-T}} \tag{36}$$

where:

T = Sample Period

$$Z = e^{ST}$$

$$G(Z) = \frac{2KZ\ (1 - e^{-T})}{(Z - 1)(Z - e^{-T})}$$

$$1 + G(Z) = \frac{(Z - 1)(Z - e^{-T}) + 2\ KZ\ (1 - e^{-T})}{(Z - 1)(Z - e^{-T})}$$

---

11 Ibid., pp. 522-524.

The closed loop system is stable if the numerator of 1 + G(X) possesses no zeros outside the unit circle in the Z plane.

$$P(Z) = 1 + G(Z) = Z^2 + \left|2K\ (1-e^{-T}) - e^{-T} - 1\right| Z + e^{-T} \qquad (37)$$

The necessary and sufficient conditions for P(Z) to have no zeros outside the unit circle are:

$$|P(0)| < 1,$$
$$|P(1)| > 0$$
$$|P(-17)| > 0.$$

Applying these three conditions,

$$P(0) = e^{-T} < 1 \qquad (38)$$

$$P(1) = 2K(1 - e^{-T}) > 0 \qquad (39)$$

$$P(-1) = 2 - 2K(1 - e^{-T}) + 2e^{-T} > 0 \qquad (40)$$

The first two conditions are satisfied for any positive gain and the third condition places an upper bound on K.

$$T_1 2K(1 - e^{-T}) < + 2(e^{-T} + 1) \qquad (41)$$

$$K < \frac{1 + e^{-T}}{1 - e^{-T}}$$

The sampling period (T) is 3 seconds, therefore, for the indicator system to be stable:

$$K < \frac{1 + e^{-3}}{1 - e^{-3}} = \frac{1 + \frac{1}{e^3}}{1 - \frac{1}{e^3}} \qquad (42)$$

$$= \frac{1 + \frac{1}{20}}{1 - \frac{1}{20}}$$

$$= \frac{1.05}{.95}$$

$$K < 1.1$$

The K of the RADUX indicator system is approximately 0.0136 so the system is extremely stable.

3.8.2 Motors

The motors used in the phase shifter indicator unit are size-15, 60-cps, 2-field servo motors. They are integrally mounted on gearhead reducers. The largest speed reduction practical was taken in the gearhead, which in this case is 77:1.

3.8.3 Gear Ratio Calculations

3.8.3.1 Servo Motor To Resolver, Low Speed

The reduction from the motor to the resolver is 399,259:1. Assuming a motor speed of 2,500 revolutions per minute, the resolver speed in the unit of measurement would be

$$\frac{2{,}500\ \text{rev/min}}{399{,}259} = 0.00625\ \text{rev/min} \tag{43}$$

If one resolver revolution requires 5,000 microseconds, the resolver speed in microseconds per minute becomes

$$(5{,}000\ \mu\text{sec/rev})(0.00625\ \text{rev/min}) = 31.25\ \mu\text{sec/min} \tag{44}$$

And if furthermore one microsecond is approximately equal to 0.162 knots, the tracking rate is

$$(31.25\ \mu\text{sec/min})(0.162\ \text{knots}/\mu\text{sec})(60\ \text{min/hr}) = 304\ \text{knots} \tag{45}$$

the time required to reduce a 200-microsecond error in low speed is approximately

$$\frac{200\ \mu\text{sec}}{31.25\ \mu\text{sec/min}} = 6.4\ \text{min} \tag{46}$$

3.8.3.2 Servomotor To Resolver, High Speed

The reduction from the motor to the resolver is 24,954:1 in the high-speed condition. This rate is 1/16 the reduction in low speed, therefore the resolver speed in revolutions per minute is

$$(16)(0.00625\ \text{rev/min}) = 0.1\ \text{rev/min}. \tag{47}$$

The resolver speed in microseconds per minute is

$$(16)(31.25\ \mu\text{sec/min}) = 500\ \mu\text{sec/min} \tag{48}$$

A 2,500-microsecond phase difference will be reduced to 200 microseconds in approximately 4.6 minutes, i.e.,

$$\frac{2{,}300\ \mu\text{sec}}{500\ \mu\text{sec/min}} = 4.6\ \text{min} \tag{49}$$

3.8.3.3 Resolver To Counter

The reduction from the counter to the resolver is 500:1 with the other input to the differential fixed. The resolver completes one revolution in 5,000 microseconds while the counter takes 10 microseconds.

3.8.3.4 Counter To Recorder Synchro Transmitter

There is a 40:1 reduction between the counter and the recorder synchro transmitter with one transmitter revolution requiring 400 microseconds.

3.8.3.5 Counter To Remote Indicator Synchro Transmitter

The ratio between the counter and the remote indicator synchro transmitter is 1:1 with one transmitter revolution requiring 10 microseconds.

3.8.3.6 Resolver To Local Oscillator Adjusting Capacitor

The reduction from the resolver to the adjusting capacitor is 1.875:1. The capacitor may be adjusted by any one of the three resolver channels by magnetically clutching in the proper resolver as shown in Figure 17. It may also be adjusted manually from

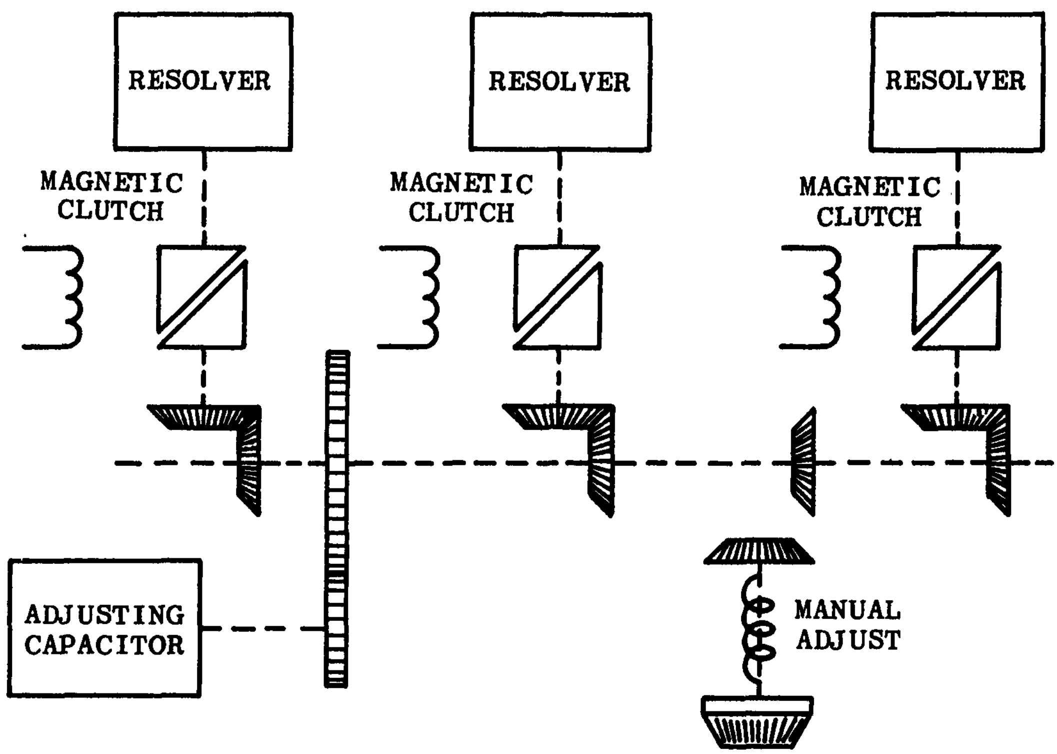


Figure 17. Capacitor Adjustment Methods

the front panel. Pushing in on the manual adjust control engages the control with the adjusting capacitor driving shaft while at the same time tripping a microswtich that de-energizes all the magnetic clutches.

3.8.4 Mechanical Design Considerations

Becaυse the indicator gearing assembly is so complex, maintainability was given prime consideration along with accuracy and reliability. The package consists of three identical gear assemblies mounted as plug-in packages on a frame containing additional common gearing. Since the three gear assemblies are identical, a minimum number of spare parts is required. To facilitate maintenance, the entire assembly may be easily removed.

The indicator unit is adequately ruggedized to withstand the vibration and shock expected in Naval service without the gears binding or losing accuracy.

3.8.5 Two-Speed Servo System

A two-speed servo system is required to provide rapid rotation of the resolvers during large error periods (error larger than 200 microseconds) and slow rotation during operation with relatively small errors. The specification requires the high-speed to reduce a 2500-microsecond phase difference to 200 microseconds within 5 minutes. This set the minimum slewing rate at 460 microseconds per minute. The specification also requires the low speed to track a 40-knot rate, which is 4.1 microseconds per minute, while requiring less than 15 minutes total receiver synchronization and phase matching time. It was decided that to meet these requirements, the low-speed must reduce a 200-microsecond phase difference to zero in less than half this period of time, which is a rate of greater than 26.7 microseconds per minute.

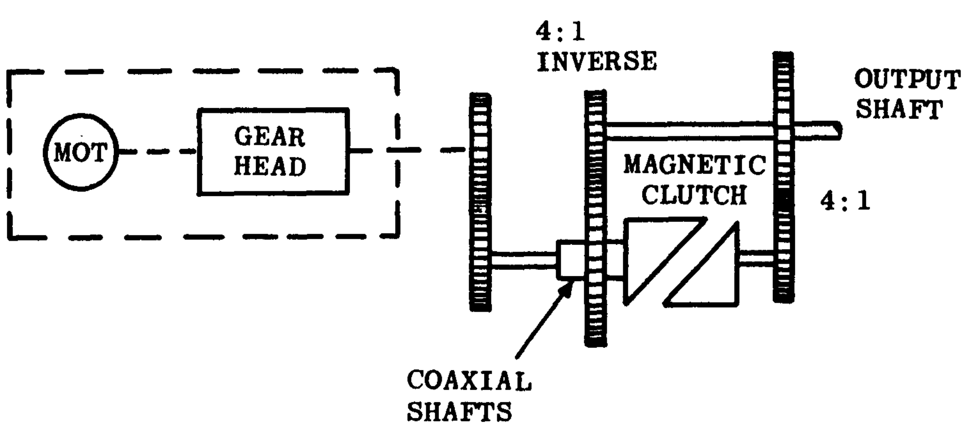


Figure 18. Two-Speed Servo System Arrangement

The present system provides a low speed of approximately 31.25 microseconds per minute and a high speed of approximately 500 microseconds per minute. The two speeds were derived as shown in Figure 18.

The servo motor drives the input to a magnetic clutch with two output shafts. In high-speed operation, one end of the clutch engages to provide a 4:1 speed increase. In low-speed operation, the other end of the clutch engages to provide a 4:1 speed reduction. This arrangement allows the high speed to be 16 times greater than the low speed.

3.8.6 Station Selection

The receiver is required to select three of four possible received signals and to indicate the phase difference between them. The phase difference must be indicated as station 1 minus station 2, station 2 minus station 3 and station 1 minus station 3.

Figure 19 is a table showing the method of selecting and displaying station assignments for each line of position. The four stations are designated A, B, C and D. A front panel switch selects any one of the four possible combinations of stations. Proper signals are switched to the resolvers and an indicator below each counter displays which two stations are assigned to each line of position. To achieve the proper readout, the counter on line of position (L.O.P.) number 1 is reversed from the counters of L.O.P's 2 and 3. The resolver of channel 2 is connected to produce an increasing phase angle for the same direction of rotation that would produce a decreasing phase angle on channel 1.

3.9 TEST RESULTS

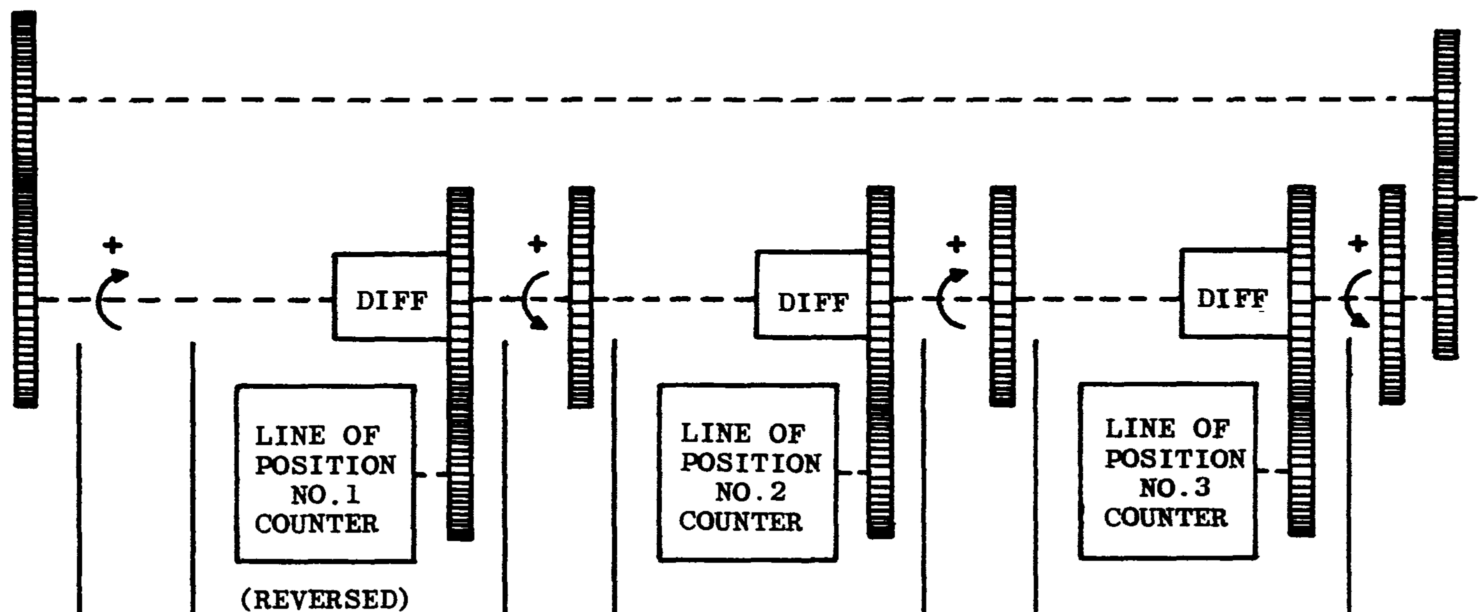
3.9.1 Receiver Selectivity and Dynamic Range

The receiver selectivity requirements, have been met in their entirety. Dynamic range design objectives have been substantially met. Tests on the final units show an expected differential phase delay of approximately 15 microseconds over a 100-db amplitude range, as compared to the design objective of 5 microseconds. Tests on the breadboard receivers have been made with differential delays of 5 or less.

3.9.2 Loop Antenna Coupler

3.9.2.1 Sensitivity

The loop antenna coupler sensitivity requirements have been met in their entirety by using antenna tuning. Sensitivity is 150 microvolts with a single loop planar and 200 microvolts with crossed loops (omni) and the antenna coupler.



STA	CHAN 1 INPUT	L.O.P. NO. 1	CHAN 3 INPUT	L.O.P. NO. 2	CHAN 2 INPUT	L.O.P. NO. 3	CHAN 1 INPUT
ABC	A	A–B	B	B–C	C	A–C	A
BCD	B	B–C	C	C–D	D	A–C	B
ACD	A	A–C	C	C–D	D	A–D	A
ABD	A	A–B	B	B–D	D	A–D	A

Figure 19. Method of Selecting And Displaying Station Assignments For Each Line Of Position

3.9.2.2 Omnidirectional Pattern

The differential phase shift with antenna rotation design objectives have essentially been met by using a unique feedback technique in conjunction with the tuned antennas. The differential phase delay measures between 2 and 5 microseconds, although theoretically, it should be much less than one microsecond. These values are attributed to the inability to adjust the r-c coupling network perfectly with existing adjustment and measurement techniques and to receiver long time stability.

3.9.3 Standard Deviation

The standard deviation design objective of 10 microseconds essentially has been met with standard deviation figures of between 10 and 15 microseconds. This result is the direct consequence of the indicator bandwidth.

3.9.4 Long-Time Stability

Long-time stability requires the standard deviation over a 24-hour period to be 5 microseconds or less with a 1-microvolt input. Test results show the requirements to be essentially met with measured standard deviations of between 7 and 13 microseconds under the required conditions.

3.9.5 Transient Response

The transient response requirements have been met with a measured time of 50 seconds to the 70-percent recovery points as compared to the 60-second maximum required by specification. The transient response is determined when the bandwidth is fixed. Slowing the response narrows the bandwidth, which in turn reduces the standard deviations under poor signal-to-noise-ratio conditions.

3.9.6 Shift In Mean

The mean reading under a poor signal-to-noise ratio is a function of the discriminator balance as well as the balance of all noise energy above and below the discriminator center frequency. The shift-in-mean requirements were met by using a trial-and-error method to adjust the shape of the receiver band-pass and discriminator balance until the shift was low enough to meet the specification. This is not adequate because a repeatable method for making the necessary adjustments is lacking. A better balancing technique will be needed for future models.

3.9.7 Tracking Lag

Tracking-lag performance tests show that the specification essentially has been met. Results show a lag of approximately 15 microseconds compared to the design goal of 10 microseconds. The lag is determined by indicator bandwidth.

3.9.8 Synchronization And Readout

The requirements of a 3-minute synchronization time and a readout within 15 minutes from an unsynchronized condition have been met. The goal of synchronizing with two stations with 1-to-5 signal-to-noise ratios has not and cannot be met with the techniques used in this equipment. Synchronizing with 4 stations under 1-to-3 signal-to-noise-ratio conditions has been accomplished with a high degree of probability.

4. CONCLUSIONS

From the test results it can be seen that the equipment delivered by Motorola under this contract substantially meets the performance specifications.

An invention has made possible the advance of the state-of-the-art to the point where the specification sensitivity requirements were met together with the requirement that the measured phase of a signal be independent of the loop antenna orientation.

PART II

RECOMMENDATIONS

5. RECOMMENDATIONS

The equipment developed under this contract is believed to be adequate electrically for performing the tasks required of it. A few exceptions, however, are discussed in the following paragraphs.

5.1 MECHANICAL DESIGN

A mechanical redesign of the receiver would be desirable to minimize interference caused by layout deficiencies. A final design, to be produced in large quantities, should be made with power-consumption and size reductions as prime objectives.

5.2 DISCRIMINATOR DESIGN

A new discriminator design would be required with a means of noise balancing as an inherent part of the circuit. If possible, an automatic arrangement should be used.